VOLUME 20 FEBRUARY, 1932 NUMBER 2

PROCEEDINGS
of
The Institute of Radio Engineers

A Message Relative to Employment
Will be Found on Page 198

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Institute of Radio Engineers Forthcoming Meetings

TWENTIETH ANNIVERSARY CONVENTION
Pittsburgh, Penna.
April 7, 8, 9, 1932

CHICAGO SECTION
February 11, 1932

CINCINNATI SECTION
February 16, 1932
March 15, 1932

DETROIT SECTION
February, 16, 1932
March 15, 1932

NEW YORK MEETINGS
February 3, 1932
March 2, 1932

Some of Those Present at "Old Timers" Night Held by the San Francisco Section on November 19, 1931

Standing, left to right: Robert Cannon, F. E. Dunklee, L. T. Franklin, R. A. Stack, H. E. Williams, H. E. Coyle, J. O. Watkins, L. F. Fuller, A. R. Rice.

Seated: B. H. Linden, R. M. Heintz, George O'Hara, C. A. Lindh, R. J. Robinson, W. W. Hanscom.

INSTITUTE NEWS AND RADIO NOTES

Annual Meeting of the Board of Directors

The annual meeting of the Board of Directors was held at the Institute office on Wednesday, January 6, 1932. Those present were R. H. Manson, president; Melville Eastham, treasurer; Alfred N. Goldsmith, editor; O. H. Caldwell, Lloyd Espenschied, J. V. L. Hogan, H. W. Houck, L. M. Hull, C. M. Jansky, Jr., R. H. Marriott, E. L. Nelson, A. F. Van Dyck, and H. P. Westman, secretary.

The minutes of the previous meeting were amended to include on the Tellers Committee, Beverly Dudley and the Secretary.

H. H. Gleason and W. R. Jones were transferred to the grade of Member, and M. F. Eddy was admitted to the grade of Member. Eighty-seven Associates and six Juniors were elected.

The Tellers Committee submitted its report on the count of ballots cast. Those candidates receiving a plurality of votes case were declared elected, and are as follows: President, W. G. Cady; Vice President, E. V. Appleton; Directors, O. H. Caldwell and E. L. Nelson.

Melville Eastham was reappointed Treasurer for 1932 and H. P. Westman was reappointed Secretary for the year.

In accordance with the new Constitution, five Directors were appointed by the Board, and are Arthur Batcheller, Alfred N. Goldsmith, H. W. Houck, C. M. Jansky, Jr., and William Wilson.

Section 63 of the By-Laws to the Constitution was amended to read as follows:

> "The Secretary is authorized to receive annual subscriptions to the monthly PROCEEDINGS at the rate of $10.00 per annum with an extra postage charge sufficient to cover the mailing cost to all countries to which the bulk rate of postage does not apply. A discount of fifty per cent from the subscription price of $10.00 will be allowed to colleges, public libraries, and libraries of learned organizations or institutions of standing, upon direct subscription to Institute headquarters. Members, publishers and subscription agencies may be allowed a discount of twenty-five per cent."

It was agreed that the Institute should be represented at a meeting which will consider the advisability of establishing a Sectional Committee on Noise Measurement under American Standards Association procedure. The appointment of this representative was left to J. W. Horton, Chairman of the Standards Committee.

A report of the Constitution and Laws Committee on its first review of the legislation relating to radio before Congress was considered, and the Committee instructed to continue with its deliberations.

In view of the widespread unemployment among the members of the Institute, an Emergency Employment Committee was established. Further details of it appear at the end of this particular report.

It was announced that the paid membership as of December 1, 1931 was 6,734, compared with 6,535 as of December 31, 1930 This is a net gain in membership of 199 members during 1931.

Emergency Employment Committee

An Emergency Employment Committee was established by the Board of Directors at its January 6th meeting in order that the radio field may be surveyed to determine how many members of the Institute and other radio engineers are unemployed. In order to assist those in need and keep them in food and shelter, the committee was directed to determine what new work could be started and what funds could be obtained to support it.

The committee, under the chairmanship of R. H. Marriott, immediately started operations and has already circularized the membership requesting those who are in need to bring their cases to its attention, and also soliciting funds from those who are able to contribute to the assistance of their fellow members.

Records of the qualifications of those out of employment are being collected, and when the industry needs men these will be available to permit suitable candidates to be located with minimum trouble.

A general survey of broadcast station coverage is planned as a type of work in which these members can be employed at nominal salaries and which work will not interfere with those at the present time employed in the radio industry. It is anticipated that this will help to support them until the industry has suitable occupations for them.

The committee is keenly interested in receiving comments from the membership and suggestions as to methods of improving conditions existent in the radio employment field. Those who are interested in participating and contributing to the broadcast survey, which is the first of the plans which will be put into operation, are urged to forward their suggestions and contributions to Institute headquarters immediately.

To be effective the committee's plans must be put into operation with a minimum delay and all members are urged to participate at the earliest possible moment.

Radio Transmissions of Standard Frequency

The Bureau of Standards transmits standard frequencies from its station WWV, Washington, D.C., every Tuesday. The transmissions are by continuous-wave telegraphy at 5000 kilocycles, and are given continuously from 2 to 4 P.M., and from 8 to 10 P.M., Eastern Standard Time. This service may be used by transmitting stations in adjusting their transmitters to exact frequency, and by the public in calibrating frequency standards and transmitting and receiving apparatus. They can be heard and utilized by stations equipped for continuous-wave reception throughout the United States, although not with certainty in some places. The accuracy of the frequency is at all times better than a part in a million.

The transmissions consist mainly of continuous, unkeyed carrier frequency, giving a continuous whistle in the phones when received with an oscillatory receiving set. The first five minutes of the transmission consist of the general call (CQ de WWV) and announcement of the frequency. The frequency and the call letters of the station (WWV) are given every ten minutes thereafter.

From the 5000 kilocycles any frequency may be checked by the method of harmonics. Information on how to receive and utilize the signals is given in Bureau of Standards Letter Circular 314, a copy of which may be obtained on request addressed to the Bureau of Standards, Washington, D.C.

The Bureau is desirous of receiving reports on these transmissions, especially because radio transmission phenomena change with the season of the year. The data desired are approximate field intensity, fading, and the suitability of the transmissions for frequency measurements. It is suggested that in reporting upon field intensities for these transmissions, the following designations be used where field intensity measurement apparatus is not at hand; (1) hardly perceptible, unreadable; (2) weak, readable now and then; (3) fairly good, readable with difficulty; (4) good, readable; (5) very good, perfectly readable. A statement as to whether fading is present or not is desired, and if so, its characteristics, such as whether slow or rapid, and time between peaks of signal intensity. Statements as to type of receiving set used in reporting on the transmissions and the type of antenna used are likewise desired. The Bureau would also appreciate reports on the use of the transmissions for purposes of frequency measurement or control.

All reports and letters regarding the transmissions should be addressed to be Bureau of Standards, Washington, D.C.

Proceedings Binders

Binders for the PROCEEDINGS, which may be used as permanent covers or for temporary transfer purposes, are available from the Institute office. These binders are handsome Spanish grain fabrikoid, in blue and gold. Wire fasteners hold each copy in place and permit removal of any issue from the binder in a few seconds. All issues lie flat when the binder is open. Each binder will accommodate a full year's supply of the PROCEEDINGS and they are available at one dollar and seventy five cents ($1.75) each. Your name, or PROCEEDINGS volume number, will be stamped in gold for fifty cents (50c) additional.

Bound Volumes

The twelve issues of the PROCEEDINGS published during 1930 are now available in blue buckram binding to members of the Institute at nine dollars and fifty cents ($9.50) per volume. The price to nonmembers of the Institute is twelve dollars ($12.00) per volume.

1931 Index to the Proceedings

The 1931 Index to the PROCEEDINGS was issued as a supplement to the January, 1932, issue. The Institute will be glad to mail extra copies upon request.

Committee Work

ADMISSIONS COMMITTEE

A meeting of the Admissions Committee was held at 10 A.M. on Wednesday, January 6th, at the office of the Institute and was attended by C. M. Jansky, Jr., chairman; C. N. Anderson, H. C. Gawler, R. A. Heising, A. V. Loughren, and H. P. Westman, secretary. Three applications for transfer to the grade of Member were approved and three applications for admission to the grade of Member were denied.

CONSTITUTION AND LAWS COMMITTEE

A meeting of the Constitution and Laws Committee was held on Wednesday, December 16th, at the Institute office. Those present were R. H. Marriott, chairman, A. B. Chamberlain (representing E. K. Cohan), Lloyd Espenscheid, W. G. H. Finch, H. E. Hallborg, H. C. Gawler, R. A. Heising, L. C. F. Horle, R. H. Langley, and H. P. Westman, secretary.

The committee reviewed a number of laws pending in Congress, affecting radio and attempted to formulate some general policies under which it might operate in a more extended consideration of these laws than it was able to give the subject in a single meeting.

Membership Committee

The Membership Committee held its regular monthly meeting at the Institute office on Wednesday, January 6, 1932, the following being in attendance: H. C. Gawler, chairman; C. R. Rowe, J. E. Smith and A. M. Trogner.

Tellers Committee

The Tellers Committee met at 1:30 P.M. on Wednesday, January 6th, and prepared its report on the count of ballots received in the election of officers for 1932. Those present were B. Dudley, H. W. Houck, A. F. Van Dyck, and H. P. Westman, secretary.

STANDARDIZATION

Executive Committee of the Standards Committee—I.R.E.

A meeting of the Executive Committee of the Institute's Standards Committee, which is composed of the Chairmen of the Technical Committees and the Subcommittees, was held at 10:15 A.M. on Wednesday, December 16, 1931, in the office of the Institute. Those attending were J. W. Horton, chairman; E. D. Cook, F. H. Engel, J. V. L. Hogan, B. E. Shackelford, H. M. Turner, H. A. Wheeler, William Wilson, H. P. Westman, I.R.E. Secretary, and B. Dudley, Standards Committee Secretary.

Six Technical Committees have been appointed to review the report of the previous Committee on Standardization and to add to this material such matters as are at the present time suitable for standardization. These committees are as follows: Fundamental Units and Measurements, H. M. Turner, chairman; Radio Receivers, H. A. Wheeler, chairman; Transmitters and Antennas, W. Wilson, chairman; Vacuum Tubes, B. E. Shackelford, chairman; Electro-Acoustic Devices, E. D. Cook, chairman; and Electro-Visual Devices, J. V. L. Hogan, chairman.

A detailed discussion of the work of these various technical committees was held and methods of correlating their work established. The necessity of a close coöperation between the committees considering allied subjects was pointed out. Various editorial considerations were established so that the reports of these six committees might readily be combined in the final report without any necessity for extensive editing at that time. A time schedule was prepared looking forward to having the final report of the Standards Committee in finished form for submission to the Board of Directors for final approval by December 1, 1932.

A number of problems which are of particular importance and prominence in the field of radio standardization were outlined and submitted to the proper technical committees for early consideration.

Technical Committee on Radio Receivers—I.R.E.

The Technical Committee on Radio Receivers of the Institute held a meeting on Thursday, January 7, 1932, at the Institute office. The meeting was attended by H. A. Wheeler, chairman; A. F. Barden (representing David Grimes), C. M. Burrill, E. T. Dickey, V. M. Graham, F. A. Hinners, R. H. Langley, H. O. Peterson (representing H. H. Beverage), A. E. Thiessen, E. W. Wilby (representing David Grimes), and B. Dudley, secretary.

As this was the first meeting of the Technical Committee this year, a thorough discussion was made of the material to be considered by the committee and some general policies for its operation were established. A considerable discussion was held regarding present-used methods of broadcast receiver measurement and some material on the subject which had previously been adopted as standards.

Technical Committee on Vacuum Tubes—I.R.E.

A meeting of the Technical Committee on Vacuum Tubes of the Institute, was held at 10:15 A.M. on Tuesday, December 15th, at the Institute office and was attended by B. E. Shackelford, chairman; N. P. Case, J. F. Hanley, A. Lederer, E. E. Spitzer, Dayton Ulrey, J. C. Warner, K. S. Weaver, P. T. Weeks, and B. Dudley, secretary.

The committee considered some recommended changes in letter symbols used in vacuum tube work and as the matter was rather involved, it was decided that it be placed in the hands of the Subcommittee on Symbols, operating under the technical committee.

The report on amplifier classifications published in the last report was considered and some recommendations regarding changes in this classification were made. The committee then reviewed a number of definitions appearing in the 1931 report and established some changes which it is believed will be desirable when these definitions are approved by the 1932 committee.

Institute Meetings

Atlanta Section

The November meeting of the Atlanta Section was held on the 30th at the Atlanta Athletic Club and Ansley Hotel, Atlanta, and was presided over by H. F. Dobbs, chairman.

The paper of the evening, "With Byrd in the Antarctic", was presented by Lieutenant Malcolm P. Hanson, U.S.N.R. Lieutenant Hanson's lecture was well illustrated by many lantern slides showing details of the trip made by Commander Byrd to the South Pole, and the interest of those in attendance was such as to continue the general discussion even after the formal closing of the meeting.

The Atlanta Section was particularly fortunate in having as its guests the Atlanta Sections of the American Institute of Electrical Engineers, the American Society of Mechanical Engineers, and the Atlanta Radio Dealers Association. The attendance at the meeting totaled one hundred and fifty, twelve members and guests attending the informal dinner which preceded it.

Cincinnati Section

A meeting of the Cincinnati Section was held on December 15th at the Engineers Club in Dayton, Ohio, Acting Secretary-Treasurer H. G. Boyle, presiding.

A paper by W. L. Everitt, Professor of Electrical Engineering, Ohio State University on "Theoretical Analysis of Class B and C Amplifiers" was presented.

The author gave a general consideration of type "A" amplifiers designating their limitations and the distortion normally present and the usual causes of it. The operation of push-pull systems and their characteristics was also pointed out. Class "B" amplification was discussed as was the usual type of radio-frequency amplifiers and modulators.

The use of class "C" amplifiers as modulators was covered. All of the various types of amplifiers were illustrated by theoretical and actual plate-current curves so that some idea of the distortion to be expected in the various classes could be had.

The discussion was participated in by Messrs. Boyle, Felix, Kilgour, and Osterbrok of the sixty-two members and guests who attended the meeting.

Cleveland Section

On December 18 the annual meeting of the Cleveland Section was held at Case School of Applied Science in Cleveland, G. B. Hammon, presiding.

A paper on "Considerations in Air Cell Receiver Design" by F. T. Bowditch was presented. As this paper appears elsewhere in this issue of the Proceedings it will not be necessary to summarize it here.

The yearly election of officers was held and the following elected for 1932: Chairman, E. L. Gove, Radio Air Service Corporation; Vice

Chairman, P. A. Marsal, National Carbon Company; and Secretary-Treasurer, Deane Kintner, Radio Editor, *The Plain Dealer*.

The meeting was attended by forty-three members and guests.

Connecticut Valley Section

The November 19 meeting of the Connecticut Valley Section was held at the Hartford Electric Light Company auditorium, R. S. Kruse, chairman, presiding.

Members of the Hartford Engineers Club and the local section of the American Society of Steel Treaters were invited to attend and hear a paper on "X-Ray Analysis of Metals," by H. H. Lester, Research Physicist, Ordinance Department, Watertown Arsenal, Watertown, Mass.

Dr. Lester's paper covered the use of X-rays in locating defects in steel and also treated the subject of analysis of the atomic structure of steel by means of X-Rays and the microscope.

Several of the forty-four members and guests in attendance entered into the discussion of the paper.

The annual meeting of the Connecticut Valley Section was held on December 17 at the Hotel Charles, Springfield, Mass., R. S. Kruse, chairman, presiding.

The election of officers for 1932 was held with the following results: Chairman, L. F. Curtis, United American Bosch Corporation, Springfield, Mass.; Vice Chairman, H. W. Holt, Westinghouse Electric and Manufacturing Company, Chicopee Falls, Mass.; Secretary-Treasurer, George Grammer, American Radio Relay League, West Hartford, Conn. Mr. Curtis then took the chair.

A paper on "Radio-Frequency Pentodes" was presented by E. W. Ritter of the RCA Radiotron Corporation of Harrison, N. J.

The author covered the development of the new type 239 radio-frequency pentode and the discussion which followed the paper was participated in by a large number of the thirty-two members and guests in attendance.

Detroit Section

The annual meeting of the Detroit Section was held on December 18 in the Detroit News Conference Room, L. N. Holland, presiding.

A paper on "The Operation of a Modern Police Radio System" was presented by E. C. Denstaedt, Supervisor of Police Radio of Detroit, Mich.

The speaker outlined the growth of the idea (which originated with the Detroit Police Department in 1921), giving the history of the experiments which resulted in the system as used today. A description

of the early transmitters and receivers followed. Although a measure of success was achieved in the early (1921–1926) experiments, results were not entirely satisfactory, and it was not until 1928 that successful short-wave work was made possible by the development of screen-grid receiver tubes. The speaker pointed out the necessity of the transmission of police calls being outside of the broadcast band of frequencies.

A detailed description of the transmitters now in use was made. It was pointed out that in most large cities transmitters had to be located in various parts of the city to insure full coverage since the Federal Radio Commission limited the output of Police Radio Stations to 500 watts. At the transmitter in Belle Isle, Detroit, all lines and equipment are duplicated to insure uninterrupted operation.

The speaker then described the sensitivity and type of receivers now in use. Data were also given regarding the cost of operation and maintenance of the receivers.

The paper was discussed by a number of the fifty-seven members and guests who attended the meeting.

To make it a complete Police Department night, three very fine entertainers were furnished through the courtesy of the Detroit Police Department. Their program was enjoyed by every one present.

Los Angeles Section

The December meeting of the Los Angeles Section was held on the 15th of the month at the Mayfair Hotel, Chairman T. E. Nikirk, presiding.

Two papers were presented, the first, "Radio and Communication Facilities of the Airways Division of the U. S. Department of Commerce" by T. K. Johnson, Assistant Airways Traffic Supervisor of the U. S. Department of Commerce, and the second, "A Technical Description of Aural and Visual Types of Radio Beam Transmitters Now Used by the Department of Commerce" by James H. Nicholson, Radio Electrician, U. S. Department of Commerce. In addition to the presentation of these papers, a teletypewriter machine was installed in the meeting hall through the courtesy of the local telephone company and connected in on the regular weather reporting circuits in use between Pacific Coast airports.

As this was the annual meeting, reports of the various committee chairmen were presented and a brief summary of the financial conditions of the section was given by the Secretary-Treasurer. The election resulted in the following selection of officers for 1932: Chairman, Ernest H. Schreiber, Southern California Telephone Company, Los Angeles; Vice Chairman, L. Elden Smith, Security-First National

Bank of Los Angeles; and Secretary-Treasurer, H. W. Andersen, Pathe Studios, Culver City, California.

The meeting was attended by sixty members and guests of whom thirty-five were present at the informal dinner which preceded it.

Philadelphia Section

The November meeting of the Philadelphia Section was held on the 18th at the Engineers Club, G. W. Carpenter, chairman, presiding.

"Present Trends in Transmitter Developments" was the subject of the paper by J. B. Coleman, a member of the RCA Victor Company at Camden, N. J.

The author discussed the work of the U. S. Federal Radio Commission and the International Technical Consulting Committee on Radio in establishing basic requirements for commercial and broadcast transmitters. Power ratings for broadcast transmitters in this country and Europe were compared. Improvement in frequency stability, fidelity, and the reduction of background noises was pointed out and various methods of modulation compared.

Transmitters for broadcast service ranging from 250 watts to 50 kw and for police and commercial service were described. Special equipment for Hawaiian Islands service operating in the frequency range of 35.3 to 54 megacycles was also discussed.

The meeting was attended by sixty-five members and guests, a number of whom participated in the general discussion which followed the presentation of the paper.

Rochester Section

A meeting of the Rochester Section was held on December 3 at the Hotel Sagamore, George Wright, presiding.

A paper by Phillips Thomas, Research Engineer of the Westinghouse Electric and Manufacturing Company, on "Electrons at Work and Play" was presented.

Dr. Thomas demonstrated many new developments in science, among which were the Strobglow, an apparatus which makes it possible to examine any kind of rotating or vibrating machinery as though the piece under examination were stationary; the breath relay, an application of the phototube or "Electric Eye" by means of which it is possible to blow out all the lights of a room and to light them again with a match. The modern William Tell, the "photomatic telephone," the fire "scanner," visible sound, synthetic pictures, and many other demonstrations that in olden times would have been regarded as "black magic".

The attendance at the meeting totaled three hundred and ten.

Each month the Rochester Engineering Society, of which the Rochester Section of the Institute is a member, holds a series of luncheons and those held during the month of December were sponsored by the Institute Section. Four meetings were held and the following speakers discussed the subjects enumerated:

December 8—Ray H. Manson, Vice President and Chief Engineer of the Stromberg-Carlson Telephone Manufacturing Company, spoke on "The Expanding Field of Radio."

December 15—Rowland G. Edwards, Director, Auditorium Permanent Players, discussed "The Showman in Business."

December 22—George E. Norton, Rector, St. Paul's Episcopal Church, presented "Science and Poetry at Christmas."

December 29—John F. Ancona, Consulting Engineer, gave a paper on "The Education of an Engineer."

San Francisco Section

The November meeting of the San Francisco Section was held at the Stewart Hotel, San Francisco, on November 18, with fifty-eight members and guests present. The speaker of the evening was L. F. Fuller who spoke on "Landmarks in Radio Development." Dr. Fuller divided these "landmarks" into primary and secondary orders, the first including the fundamental principles which have remained unchanged through the advance of the art, while the secondary landmarks, although marking advances at the time of their discovery and having a profound influence on development, have been left behind or lost sight of in the march of progress.

The meeting was characterized by the large number of "old-timers" present, fifteen, or more than a quarter of the entire attendance, having been actively engaged in radio work since 1912 or before.

An exhibit of old-time equipment was shown, some of this equipment being visible in the frontispiece to this issue. Among this apparatus may be seen the first Poulsen arc brought to this country from Denmark by C. F. Ellwell, as well as one of the first and most successful of the early generators of continuous waves for radiotelephone use. Other equipment shown included a Marconi magnetic detector with its accompanying tuner, an early crystal tuner, a rather complete collection of commercial crystal and electrolytic detectors, and a collection of vacuum tubes starting with the Fleming valve and the earliest types of audion and continuing through the various stages of development to present-day types.

In the discussion following Dr. Fuller's paper each of the old-timers shown in the group was called upon for reminiscenses of the

early days of radio, some of the speakers carrying the stories of their personal experience back for nearly thirty years. The stories told included accounts of early navy and army work, when "contrary to expectations the signals were received clearly over a distance of nearly thirty miles"; of early work with commercial companies whose identity has long been lost through mergers and otherwise; and of early telephone experiences in the days when the current-carrying power of the microphone limited the output of a transmitting station. The true story was also told of the operator who did *not* resign (as has been reported) because of his constant loss of received signals through QRM generated by a blue-bottle fly buzing around the station.

The meeting was preceded by an informal dinner which was attended by aproximately half the number attending the meeting.

The December meeting of the San Francisco Section was held on the 16th at the Bellevue Hotel, Chairman R. M. Heintz, presiding.

A paper by D. I. Cone on "Carrier-Current Telegraphy and Telephony" was presented and a general discussion followed which was participated in by many of the forty-two members and guests in attendance. The informal dinner which preceded the meeting was attended by nineteen.

Seattle Section

A meeting of the Seattle Section was held on December 17th in Guggenheim Hall, University of Washington, Seattle, Abner R. Willson, chairman, presiding.

The paper of the evening "Radio in the United States Navy" was presented by Lieutenant Commander H. H. Bouson, U. S. N.

The speaker outlined with the aid of suitable illustrations the general activities of the Navy in the radio field from 1910 to date. He discussed the methods of operations used at present, giving various reasons for these particular methods, together with a preliminary outline of the development work undertaken by the Navy and the changes that these developments have made necessary in the Naval communications system.

Discussion of the paper was participated in by Messrs. Cooper, Libby, and Willson.

As this was the annual meeting, the Nominating Committee presented its report and the following officers were elected for 1932: Chairman, Leslie C. Austin, Woolley and Company; Vice Chairman, Howard F. Mason, Washington Technical Institute; Secretary-Treasurer, Herbert H. Bouson, U. S. Navy.

The meeting was attended by thirty-five members and guests.

Toronto Section

A meeting of the Toronto Section was held on December 9 at the University of Toronto, F. K. Dalton, presiding.

A paper by C. A. Lowery of the DeForest-Crosley Radio Corporation was presented on "Some Modern Methods Applied to the Inspection of R-F Inductors and Gang Capacitors."

The speaker gave a clear and readily understandable description of the limits, theoretical and practical, of the various coils and condensers which go to make up the radio-frequency circuits of the modern broadcast receiver. A complete a-c operated beat-frequency oscillator as used in the DeForest Radio Corporation plant in Toronto was operated to demonstrate various points of the lecture.

The paper was discussed by Messrs. Andre, Baldwin, Fox, Meredith, Patience, Pipe, and Smith of the seventy-seven members and guests in attendance.

Washington Section

The December 17th meeting of the Washington Section was held at the Continental Hotel, L. P. Wheeler, chairman, presiding.

A paper "New Jobs for Radio Engineers; Radio Today—But What Tomorrow?" was presented by O. H. Caldwell, Editor of *Electronics.*

The speaker outlined the history of previous business depressions through which this country has passed since 1854 showing that the present one follows the general tendencies of those which have preceded it. The fluctuation in sales of radio receivers and tubes was indicated by means of suitable graphs.

Uses of thermionics, other than radio, were pointed out to include to a great extent photo-electric devices. The first of these was the original selenium cell which changes its resistance under illumination, now refined and sealed in a vacuum tube; then the alkali metal tube emitting electrons, and finally the photo alkali or chemical cells in which an e.m.f. is generated by the action of light. Among the numerous uses mentioned were the turning on of flood lights when an automobile approaches, the opening of garage doors under the control of head lamps, safety devices on punch presses, counting the number of bees going in and out of a hive to permit an estimation of the average amount of honey produced per bee, and counting the number of automobiles entering and leaving a tunnel so that when the difference becomes excessive the ventilating system may be automatically started.

The electrical production of musical tones by curves plotted on glass wheels thus causing a variation of light reaching the phototube beyond the curve to operate to give any desired wave form and hence

a note of any desired timber was described. The matching of colors in widely separated places, the timing of races, and control of illumination in school rooms, lowering the intensity of automobile head lamps at the approach of another car, taking the sun's position in cloudy weather by the utilization of infra-red light and the possibility of transmitting by television the progress of a total eclipse were considered in detail.

Other nonradio uses for thermionic tubes described were the electric cardiograph, the fever machine to raise body temperature locally, the radio knife for surgical operations, the control of steam boiler pressure by means of thyratrons, the transmission of electric power by direct current instead of alternating current, the control of electric signs, stroboscopic study of machinery and bearings in motion, and the use of radio music in factories so that its rhythm tends to overcome the disagreeable effects of factory noises.

The paper was discussed by Mr. Guthrie, Professor Robinson, Dr. Wheeler and others of the sixty-five members and guests in attendance, twenty of whom attended the informal dinner which preceded the meeting.

Personal Mention

A. A. Armer, formerly with the Magnavox Company is now an engineer for Frank Rieber, Inc., of San Francisco, Calif.

Previously with the Oxford Radio Corporation, R. W. Augustine is now connected with the Utah Radio Products Company at Chicago.

J. R. Bird, formerly graduate student at Massachusetts Institute of Technology has joined the engineering staff of Bell Telephone Laboratories.

C. V. Boyson previously with Federal Telegraph Company is now with the International Communications Laboratories, Newark, N. J.

Formerly with the Radio Corporation of America, F. R. Brick, Jr. is now with the American Radio News Corporation, New York City.

Previously Secretary-in-Chief of Tsing Hua University of Peiping, China, Yun Chu is now a technical expert for the Ministry of Communications at Nanking, China.

W. H. Connell has joined the engineering staff of the Gramophone Company, Ltd., Hillingdon, Middlesex, England, formerly being connected with the Columbia Graphophone Company of London.

J. R. Cubert has left the United Air Cleaner Company to become an engineer for Story and Clark Radio Corporation of Chicago.

Formerly with the General Electric Company, W. S. Duttera is now doing plant operation and engineering for the National Broadcasting Company, New York City.

F. E. Fisher has left the research department of the Phillips Petroleum Company to join the research and development division of the Precision Scientific Company, Chicago.

J. D. R. Freed is now chairman of the Board of Directors of the Freed Television and Radio Corporation.

Formerly with the Canadian General Electric Company at Toronto, Donald Dunn has become a radio engineer for Johnson Talking Machine Company, Ltd., of Liverpool, England.

Formerly chief development engineer of the Crosley Radio Corporation, D. D. Israel has become chief engineer of the radio division of the Grigsby-Grunow Company of Chicago.

F. D. Leslie previously with the Fox Hearst Corporation in New York City is now technical director for Europe of Movietonews, Inc., at Paris.

Rene Lince previously with the Standard Telephones and Cables, Ltd., of Hendon, England, has entered the Valve Department of Lissen, Ltd., Isleworth, Middlesex, England.

H. J. Loftis formerly of General Motors Radio Corporation has become chief engineer of the Henrite Products Corporation, Ironton, Ohio.

H. J. Love has become chief engineer of Sexton's of London, England.

Formerly associate physicist at the Bureau of Standards, C. G. McIlwraith has become senior electrical engineer of the U. S. Coast and Geodatic Survey, Washington, D. C.

Previously head of the Receiver Engineering Department of the Jenkins Television Corporation, H. G. Miller has joined the staff of Radio Inventions, Inc., of New York City.

O. B. Parker has become chief engineer of the Pacent Electric Company, New York City.

H. A. Schryver formerly chief engineer with the Continental Radio Corporation has become a radio engineer for Boydell Corporation, Ft. Wayne, Ind.

J. Steffensen previously engineer for A/S "Elektromekano" of Copenhagen, Denmark has become a radio engineer in the Radio Engineering Department of the Danish State Postal and Telegraph Service at Copenhagen.

Formerly chief engineer of Cable Supply Company, C. F. Stromeyer is now director of research of the Cable Radio Tube Corporation of Brooklyn.

Lincoln Walsh formerly development engineer for the Colonial

Radio Corporation is now doing consulting work with headquarters at Elizabeth, New Jersey.

C. E. Weaver previously research engineer for Electric-Therapy Products Corporation is now research engineer for General Laboratories of California at Los Angeles.

Previously with the Brunswick-Balke-Collender Company, R. S. Yoder has become an engineer for Lear Developments, Inc., of Chicago.

Lieutenant H. C. Behner, U.S.N., has been transferred from the U.S.S. Wright to the U.S.S. Langley.

Lieutenant L. Randall Daspit, U.S.N., has been transferred from the U.S.S. Gilmer to the U.S.S. Childs.

Formerly a radio engineer for the Standard Telephone and Cables, Ltd., M. C. Randall is now manager of the technical service department of F. C. Heayberd and Company, London, England.

Part II
TECHNICAL PAPERS

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BATTERY DESIGN PROBLEMS OF THE AIR CELL RECEIVER*

By

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Summary—*The present paper deals with those design features of battery-operated radio receivers which are important from the standpoint of obtaining the maximum useful battery life. The properties of the air cell A battery are discussed with relation to receiver design. The desirability of providing adequate performance until the B batteries have fallen to a very low voltage is shown. An analysis of several means of obtaining a satisfactory rate of grid voltage reduction with falling B battery voltage is included, together with a discussion of B battery resistance. It is shown that the considerations treated control a variation in useful B battery life of the order of 50 per cent.*

Introduction

THE past year in the radio industry has been characterized by a revival of interest in battery-powered radio receiving sets. Practically every radio manufacturer of prominence has developed a new battery receiver, and any one of these has performance characteristics thought to be economically impossible in this type of receiver a few years ago. Although sensitivity, selectivity, and quality are all three far in advance of previous standards for battery-operated sets, the new receivers are more economical to operate and require less service attention than ever before.

Two major factors in this state of affairs are the development of a new two-volt tube and a new source of A power known as the air cell battery. A complete line of tubes is presented, consisting of a screen-grid, general purpose, triode output, and pentode output tube. All have been designed particularly for battery service, giving the maximum performance consistent with economical operation. Both the filament current and the plate current have been reduced with economical battery service in mind, while the performance characteristics are surprisingly close to those of the corresponding alternating-current types. To match these improved tubes, the new air cell battery supplies a sufficient quantity of primary power to run the average receiver for at least 1000 hours, providing 600 ampere hours of service without attention other than the very occasional addition of water to each of its two cells.

* Decimal classification: R366.1×621.353. Original manuscript received by the Institute, August 10, 1931. Presented before Rochester Fall Meeting, November 9, 1931.

A Battery Design Problems

Conditions have not always been so favorable to good battery receiver design, however, as the following brief résumé of earlier conditions will indicate. Prior to the introduction of the air cell battery, the designer of battery operated receivers had been confined, as regards his A supply, to a choice of either the six-inch dry cell in various series-multiple combinations, or some form of storage battery, usually of the lead-sulphuric acid type. The former has the advantage of primary power in that complete independence of external sources of electrical energy for charging purposes is achieved. However, the cost of dry cell operation is inherently high, there is the nuisance of a large number of series and multiple connections, and, probably worst of all, the voltage discharge curve of the dry cell is very steep. This last item is of particular importance because the WD and –99 type tubes with which the dry cells were used are especially vulnerable to filament overvoltage. The consumer, provided with an A voltage 50 per cent in excess of the safe maximum at the tubes, and further provided with a volume controlling rheostat, shortened the life of his tubes by too often applying the full battery voltage, in his search for greater sensitivity and higher volume. As a result, the –99 tube, which is an excellent tube when operated at its rated voltage, has been practically retired from circulation, and the public is generally of the opinion that it is a short-lived tube.

Other than the six-inch dry cell, the six-volt lead-sulphuric acid storage battery has been until recently the only other source of A power generally used in battery receivers. Its voltage discharge curve is excellent, and the line of five-volt tubes headed by the reliable 201-A developed for use with it has made possible the design of many excellent receivers. Because of the storage battery, however, these receivers were not truly independent of external sources of electrical energy. In wired homes, the battery was ordinarily supplied with some sort of trickle charging device, and caused no particular annoyance. In the unwired homes, however, particularly in rural districts, the situation was quite different. On an average of about once each month, the consumer had to take his battery to some location where charging energy was available, either doing without his radio service for a day or two or else obtaining a second battery. This procedure was not only expensive, but it required the expenditure of a great deal of time and energy on the part of the consumer. Considering these limitations in regard to the A supply, and the overwhelming market acceptance of the alternating-current receiver wherever the necessary power was available, it is not surprising that the further development of the battery operated receiver was halted.

The comparative saturation of the alternating-current market has now drawn the attention of the manufacturer to the possibilities of a more intensive development of the rural or battery field, and the development of a more satisfactory form of A power to match the new tubes has made this development particularly attractive. How the air cell battery has answered the A power problem can best be shown by reference to Fig. 1, in which the comparative discharge

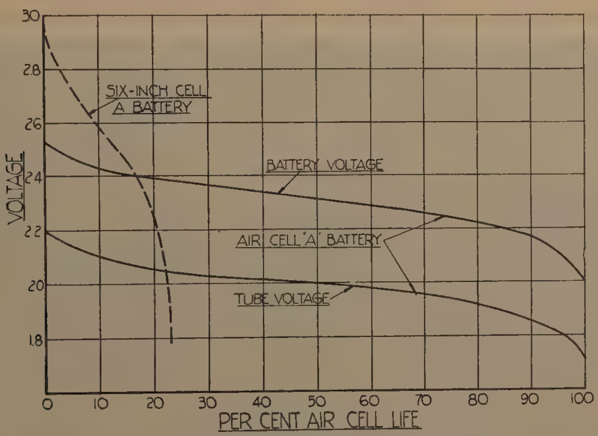

Fig. 1.—Comparative discharge curves of the air cell battery and a series-parallel group of eight six-inch dry cells, assuming an air cell receiver usage of three hours per day. Rated filament current = 0.56 amperes.

characteristics of the two forms of primary A power are shown together. In plotting this curve, a battery load of 0.56 ampere, the rated current of an average air cell receiver, is assumed, applied at the average rate of three hours per day. The six-inch dry cell curve is based on the use of eight cells in series-multiple, the most economical number, giving a battery of three volts rating. Two curves are shown for the air cell battery, the upper one being the terminal voltage of the battery, and the lower, the voltage at the tubes using a fixed series resistor. The superior performance of the air cell is self evident.

Although the discharge curve of the air cell battery leaves something to be desired from the tube manufacturer's ideal of a perfectly horizontal line, it is close enough to this ideal to permit the designer to substitute a fixed series resistor in place of the filament rheostat which has been characteristic of earlier battery receivers. If the value of this resistor is properly chosen, the entire useful energy of the battery is delivered to the tubes at usable levels, and the hazard of tube overvoltage due to careless usage of a filament rheostat is completely eliminated. If the resistor value is too low, however, the tubes may

burn out or lose emission prematurely, while if it is too high, the tube voltage will fall below a usable level before the battery is completely exhausted. The proper value for this resistance has been determined to be that which will impress a voltage of 2.2 volts at the tubes from a fresh air cell battery. From Fig. 1, it will be seen that the initial closed circuit voltage of a fresh air cell battery is 2.53 volts, which leaves 0.33 volt to be absorbed. For the conventional receiver using four screen-grid, one general purpose, and two output tubes, a series re-

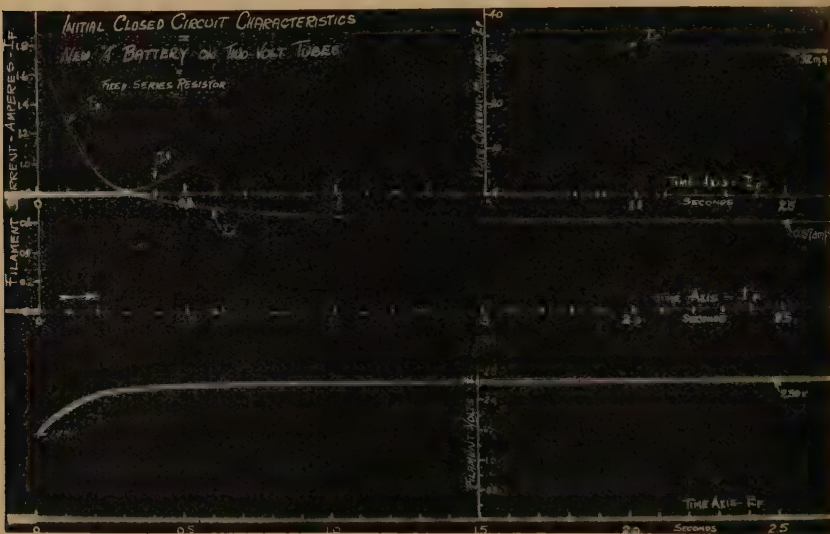

Fig. 2.—Oscillograph curves showing the initial closed circuit variation in (a) filament voltage, E_f, (b) filament current, I_f, and (c) plate current, I_p, of an air cell receiver.

sistance of 0.58 ohm is indicated. Ordinarily the lead resistance in the usual battery receiver is of the order of one-tenth of an ohm, so that it is necessary to obtain a representative average value of this resistance in any particular receiver, and subtract it from the total resistance desired in order to arrive at the proper value for the series resistor. When the resistor is properly chosen in this manner, reliable tube and battery performance throughout the life of the latter is assured, this fact having been confirmed by a large number of intermittent life tests on both tubes and batteries operated in this fashion.

While the closed circuit voltage of the air cell battery is initially 2.53 volts as stated, the open circuit voltage is ordinarily somewhat above 2.8. This excess voltage, however, is purely superficial and disappears very rapidly when a load is applied. The initial heavy surge

of current through a set of cold tubes hastens the disappearance of this voltage, while the higher voltage drop in the series resistor occasioned by this starting current serves to prevent any harm to the tubes. In less than one minute the voltage of the air cell has stabilized at a value which remains practically unchanged for the remainder of a discharge period of several hours. This phenomenon is illustrated by the oscillograph curves of Fig. 2 and the plot of Fig. 3. Fig. 2 shows simultaneous traces of filament current, filament voltage, and plate current immediately following the closing of the filament circuit of a seven-tube receiver, energized by an air cell battery through a fixed series resistor. It will be seen that the starting current is over three

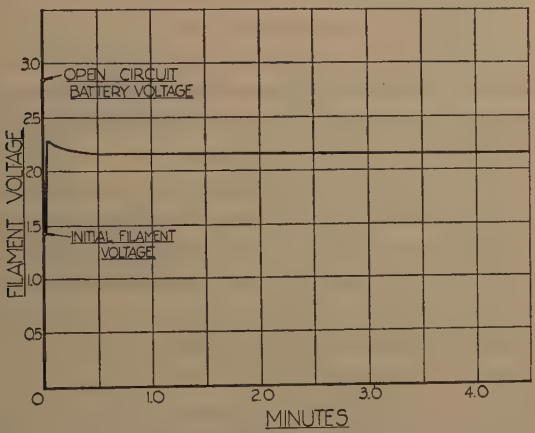

Fig. 3.—Variation in filament voltage with time following the closing of the filament circuit of an air cell receiver.

times the final stable value, while the protective action of the series resistor is clearly illustrated by the initial depression of the filament voltage. Fig. 3 shows a continuation of the filament voltage curve of Fig. 2.

The lag in filament temperature with respect to filament voltage as indicated by a comparison of the filament voltage and plate current curves of Fig. 2 is quite noticeable when a filament rheostat type of volume control is employed. In such a circuit, a fixed series resistor is used in the common filament circuit as before, while a rheostat is added in series with the filaments of the radio-frequency tubes. The novice, in operating such a volume control, will invariably turn it too far, so that when the filament temperature becomes stable at the new setting a readjustment is necessary. For this reason, some other type of volume control is ordinarily to be preferred for the air cell receiver,

such as an antenna input potentiometer, a potentiometer for the variation of screen-grid potential, control-grid potential variation, or some combination of these methods. All of these are instantaneous in their action and are to be preferred for that reason over the filament rheostat method. It should be noted that if a screen-grid potentiometer is to be employed, it is necessary to provide an added contact on the receiver switch which will remove the drain of this resistance from the B battery during idle periods. In order to keep this drain within reasonable limits during periods of receiver operation, the potentiometer should preferably have a resistance of the order of 100,000 ohms. The question of the potentiometer variation of control grid bias, involving the connection of a resistor across the C battery, will be discussed in a following section of this paper.

The air cell A battery in its present commercial form is housed in a container whose over-all dimensions are $13\frac{1}{2}$ inches long by $6\frac{3}{4}$ inches wide by $9\frac{7}{8}$ inches high, over-all. It consists of two cells, permanently connected in series, with one negative and one positive binding post for the series combination. The weight of the battery, complete with electrolyte, is $37\frac{1}{2}$ pounds. In providing cabinet space for this battery, the designer should bear in mind that the consumer must have occasional access to the battery for the purpose of inspecting the solution level in each cell and for the possible addition of water.

A comprehensive treatment of the theory of operation of the air cell battery is beyond the scope of the present paper. Briefly, however, the active components of the battery are a positive carbon and a negative zinc electrode, immersed in an electrolyte of 20 per cent caustic soda. The oxygen required for depolarization is obtained from the atmosphere through the "breather" action of the carbon electrode, being transmitted through the electrode to the electrolyte boundary in a form suited to the chemical processes of the battery. The battery is shipped dry, with the caustic soda cast in a readily soluble form around the electrodes. To put the battery in service, water only is added to each cell, any water that is suitable for drinking purposes being satisfactory. The weight of the battery as shipped, without water, is 25 pounds.

B Battery Design Problems

In the early days of battery operated receiver design, the impression prevailed that the B batteries were completely discharged when they had reached a closed circuit voltage of 17 volts per $22\frac{1}{2}$-volt section, and that nothing was to be gained by designing the receiver to give satisfactory operation at voltages below this value. Many

factors were responsible for this earlier conclusion. To begin with, the use of the soft detector, requiring the critical adjustment of a B battery tapped at every cell in the range between 16½ and 22½ volts, did not lend itself to the complete utilization of the energy in the battery to very low voltages. When the voltage of the entire battery fell below that necessary for the operation of the tube, the battery was no longer useful.

More important factors, however, are inherent to the nature of the discharge curve of the early batteries. Such a curve, plotted between hours' life as abscissas and voltage as ordinates, would be falling rapidly in voltage at 17 volts and would give but little additional life beyond that point. Battery manufacturers, in their developments to increase the high voltage service obtainable from a B battery, have flattened the discharge curve appreciably in this region, so that it requires much longer for the curve to reach 17 volts, its approach to this voltage is more gradual, and the service obtainable beyond to lower voltages is markedly increased. In the case of a typical air cell receiver now on the market, equipped with B batteries representative of modern manufacturing methods, it is estimated that approximately 50 per cent has been added to the life of these batteries by designing the receiver to operate to a B voltage of 12 volts per 22½-volt section, instead of to a 17-volt limit. The ampere hour capacity available at these lower voltages is doubly valuable because of the reduced battery drain in this region, which causes each ampere hour to last much longer than at the higher voltages.

In order to utilize this low voltage power of the B batteries, however, and thus permit the consumer to obtain the maximum service from his battery investment, it is necessary for the designer to make provision for a corresponding reduction in the various C voltages as the B voltages fall with service. Thus while a negative grid bias of 22½ volts is the proper value for a −31-type output tube when the B batteries are fresh and the plate voltage is 135 volts, it is much too high when the B voltage has fallen to 17 volts per 22½-volt section, or to a total of 102 volts, and even more so when the B voltage has fallen to 12 volts per section, or a total of 72 volts. Similarly, in the radio-frequency stages, a rapid falling off in the sensitivity of the receiver will result as the plate voltage falls if a fixed negative bias of three volts is applied to the grids of the tubes throughout the life of the B batteries. Unless some method of reducing the grid voltages to match the falling plate voltages is employed, therefore, the end of the useful life of the B batteries will be determined either by the inability of the receiver to pick up stations, or by the bad distortion of the signals that

are received. The consumer is thus compelled to waste a considerable proportion of his investment in B batteries.

Several methods are available by means of which a satisfactory rate of reduction in grid voltage with plate voltage may be accomplished. The following are typical of those now in commercial use: (a) self-bias, (b) discharge of the C battery through a resistor connected permanently across it, and (c) intermittent discharge of the C battery.

The use of self-bias, or the utilization of the voltage drop of the plate current through a resistor next the filament in the plate return circuit, is theoretically attractive. As the B battery, or, more correctly, the B plus C battery, falls in voltage, the falling plate currents automatically lower the several grid voltages. While this system has been almost universally adopted in alternating-current receiver design, its application to battery receivers has not been so generally accepted. In the first place, the fact that the cathodes or filaments of all the tubes in the battery receiver are at the same potential necessitates the common return of all the plate currents through a single resistor. This resistor is then tapped to give the several bias voltages required. Thus the unfiltered plate current variations in each tube are impressed on the grids of all the tubes in the receiver, and unusual care in filtering is required to prevent the introduction of objectionable regenerative effects. In the second place, the use of self-bias in a battery receiver is inherently more expensive from the standpoint of B battery energy. The amount of the highest grid voltage must be added to the voltage of the B battery if the tubes are to operate at the same efficiency as with a separate C battery. This added battery must have the same ampere hour capacity as the remainder of the batteries so that the self-bias system is forced to employ a B plus C battery of at least 157½ volts. The elimination of the C battery does not entirely compensate for this. Of added importance is the fact that independent control of the rate of C voltage reduction is not provided. This does not permit the designer to strike a balance between the optimum grid voltage from the standpoint of both tube operation and battery life. This phase of the problem is discussed in a later paragraph.

A reasonable compromise in the lowering of the C battery voltage may be achieved by the permanent connection of a discharge resistor across the C battery terminals within the chassis. The value of this resistor is determined by the ampere hour capacity of the C battery used, the B current demands of the receiver, the capacity of the B batteries and the estimated rate of usage of the receiver. All of these factors except the last may be evaluated with reasonable accuracy. A

satisfactory estimate of the rate of receiver usage, however, is difficult to obtain. It becomes necessary, therefore, to compromise on an intermediate rate of usage which will give every consumer some advantage over the old system of a fixed battery bias. An average usage of three hours per day has been chosen for this purpose, and resistance values are calculated with the purpose of reducing the voltage of the C battery to 16 volts per $22\frac{1}{2}$-volt section on a continuous drain during the period required for the B batteries to reach 12 volts per $22\frac{1}{2}$-volt section on the intermittent drain of the particular receiver.

The use of a 16-volt end point for the C battery is admittedly higher than the optimum value for tube operation at the corresponding plate voltage. The reason for this choice, however, is based upon the necessity of achieving a satisfactory compromise with respect to battery life. If the C voltages are reduced too fast, the corresponding increase in the total plate current as compared with that of the fixed bias system may actually result in a shorter B battery life than that obtainable under the latter system. In other words, it may require a shorter period of time for the receiver to discharge the B batteries completely under a falling bias system than would be required to discharge them partially to a higher voltage cut-off under a fixed bias system. An examination of fifteen different receiver designs showed that very satisfactory quality was obtainable throughout the entire range of B and C voltage variation under the proposed system, and that the sensitivity, while it had fallen to approximately 25 or 50 microvolts per meter, was still sufficient to bring in all local and many moderately distant stations.

A third and much more satisfactory method of achieving a proper rate of C voltage reduction is by means of a comparatively low resistance, connected to discharge the C battery during periods of receiver operation only. Its single disadvantage, as compared with the continuous discharge system just described, lies in the increased cost of the "on-off" switch of the receiver, which must now be provided with an added contact. This expense is believed justified by the increased accuracy in C voltage reduction which results, since the greatest variable of the continuous discharge system, that of the rate of receiver usage, is removed. Under the continuous discharge system, the heavy user receives only a slight benefit from a falling C voltage, since the rate of C voltage reduction is far behind the desired value. Similarly, the very light user is penalized by having the C voltages fall too fast, draining his B batteries at a rate in excess of the requirements of satisfactory operation. The use of the intermittent discharge method, on the other hand, gives every user the maximum benefit of C voltage

reduction, automatically keeping pace with the rate of usage of each particular receiver.

Another method of obtaining the proper rate of C voltage reduction, and theoretically the most accurate of all, has been developed experimentally, but the expense of its installation has not yet warranted its use commercially. The two preceding methods of C battery discharge are subject to certain variables in operation which have not been discussed heretofore. For instance, the proper value of discharge resistance is determined by the relative capacity of the B and C batteries under discussion and by the current demands of the individual receiver. Any one or more of these factors may vary sufficiently from its assumed characteristics to cause a considerable error in any one individual case. If, however, the "on-off" switch of the receiver be so constructed as to throw the C battery in parallel with a suitable portion of the B battery during the off periods, then the C battery will discharge to a voltage equal to the open circuit voltage of the former battery. This voltage, somewhat in excess of the closed circuit voltage of the B battery, is very close to the desired value of C battery voltage. Such a system, however, requires the use of at least a double pole-double throw receiver switch, and is only readily applicable when a C battery of 22½ volts or some even multiple thereof is employed, matching the taps conventionally supplied on the B batteries. It does, however, do away with all of the variables previously mentioned, leaving the variation in voltage of the chosen section of the B battery from the average of the entire battery as the chief source of error. Except in rare instances, this effect should not be of dangerous magnitude.

In view of the fact that both B batteries and vacuum tubes are now made with such great uniformity and because a considerable variation from the intended rate of C voltage reduction can be tolerated without seriously disturbing the performance of the receiver, the intermittent discharge method of C voltage reduction is believed to be the most practical. The growing practice of supplying a complete battery kit with the receiver also adds to the reliability of this method, assuring the consumer of a set of batteries definitely matched to his requirements.

B Battery Resistance

As a B battery falls in voltage on service, its internal resistance rises. At voltages higher than about 18 volts per 22½-volt section, the magnitude of this resistance is ordinarily not great enough to give concern to the designer of a battery-operated receiver. However, in order

to extend the satisfactory operation of the receiver into the lower voltage range so as to exhaust the B batteries fully as previously described, the possible effects of this resistance must be recognized in the design of the receiver. The chief source of the increased resistance with service in a dry cell is the accumulation of the waste products of the chemical reactions accompanying discharge. These gradually plug up the conducting paths within the cell and in consequence increase its resistance.

The actual magnitude of this resistance varies somewhat between batteries as to brand, previous history, and rate of discharge. Measurements have been taken on a large number of such batteries, including representative cases of the variables noted. A suitable method of measurement has been developed, which consists in reading with a vacuum tube voltmeter the alternating voltage drop produced across the battery terminals by the passage of a known alternating current through it. The conditions of measurement are thus representative of the actual behavior of the battery in a radio circuit and serve to evaluate directly the factor responsible for the regeneration phenomena about to be described. As a result of these measurements, a resistance value of 250 ohms per $22\frac{1}{2}$-volt battery section has been obtained as representing the maximum to be normally expected in radio service to 12 volts per section.

The nature of the disturbance created by the increased resistance of the B batteries has been aptly termed "motorboating." The B batteries form the common source of plate and screen-grid voltage supply for all the tubes in the receiver, and portions of their total resistance are common to any two or more of these circuits, depending upon the relative voltages applied to the circuits considered. Variations in the plate current supplied by the batteries to the output tubes, for instance, will cause a simultaneous fluctuation in the plate voltage of every tube in the receiver, due to the varying voltage drop of this current through the battery resistance. The possibilities for regeneration in such a system are very great, unless this condition has been recognized in the receiver design and suitable filtering provided. In a receiver which has not been properly designed in this respect, persistent "motorboating" of sufficient intensity to render the receiver useless for broadcast reception will usually occur when the battery resistance has risen to from 25 to 50 ohms per $22\frac{1}{2}$-volt section. The frequency of the oscillations produced in this manner is ordinarily very low, of the order of one to ten cycles per second.

In studying this condition in the laboratory, it is sufficient to investigate primarily the condition of lowest voltage and highest re-

sistance. Having corrected the design to insure satisfactory operation at this point, such operation is also insured at all higher voltages. A voltage of 12 volts per $22\frac{1}{2}$-volt section has been chosen for this purpose. Although an appreciable amount of energy usually remains in the batteries at this voltage, the sensitivity of the receiver has fallen to a point which ordinarily prohibits further usefulness. In order to improve the sensitivity, it would be necessary to lower the C voltage at a faster rate, which would simultaneously raise the B battery drains at all voltages and hasten their rate of discharge.

A battery resistance of 250 ohms per $22\frac{1}{2}$-volt section, or a total of 1500 ohms for a 135-volt battery, is normally used for this work as the result of the measurements previously described. Obviously the designer should not work too closely in the design of his filters, but should allow a reasonable factor of safety in this regard. It is good practice to provide for operation with a 25 to 50 per cent overload in the resistance factor, the added cost for this insurance against higher resistances usually being negligible.

To provide a suitable voltage supply for this investigation it is necessary either to obtain a set of B batteries which have been discharged on radio service to the voltage desired, or else to set up an artificial discharged battery by tapping fresh batteries and inserting suitable values of external resistance. It is not sufficient to reduce the voltage of fresh batteries by rapid discharge, since the resistance of a battery discharged to low voltages in this manner is not as high as that of one normally discharged over a long period of time.

In setting up an artificial discharged battery, it is important that the following points be kept in mind. First, only fresh B batteries should be employed, whose resistance is negligible. If batteries of the heavy-duty round-cell type are used, a convenient means of tapping the intermediate cells is provided by removing the back side of the carton, scraping away the sealing compound and thus exposing the bottoms of the zinc cans of the cells. Second, the resistance added must be distributed between the battery taps in proportion to the respective voltages. If all of the resistance is added on the negative end of the battery, for instance, where it is common to every B circuit, the results will not be the same as those obtained by a proportional distribution of the resistance between taps, such as occurs in actual service. Third, readings of battery voltage should be taken only under load, and should be taken across the terminals of the battery cable so as to include the voltage drop of the load current through the added resistances.

Having provided a suitable source of B power, it is next necessary to provide a source of reduced C voltage. The "$-22\frac{1}{2}$-volt" bias should

be supplied with 16 volts, and the other grid voltages should be reduced in proportion. For instance, a 3-volt bias should be reduced to 2.13 volts, and a 4½-volt bias to 3.2 volts. A potentiometer circuit should be employed to enable the accurate adjustment of these voltages.

Having such a B and C power source available, it is a relatively simple matter to determine whether the receiver under investigation will give satisfactory operation throughout the entire useful life of these batteries. If oscillation is encountered, the critical circuits may be isolated and suitable filtering provided. A convenient means of accomplishing this isolation is to provide each circuit or combination of circuits in turn with a separate source of B power, the disappearance of the oscillation indicating the isolation of the critical circuit.

It will generally be found that two or more circuits are at fault. The output tube circuit generally initiates a disturbance of the over-all B voltage, which is fed into one or more preceding circuits to cause oscillation. Sometimes it will be found less expensive to filter the output tube plate circuit and so minimize the current variations in that circuit which pass through the B batteries. On the other hand, it may prove to be cheaper to allow these currents to flow as before, and to apply filters to the one or more preceding circuits into which the feedback is least desirable. Each set design presents an individual problem in this respect, although the use of a choke coil-condenser filter in the output tube plate circuit is the most generally applicable remedy.

As to the magnitude of the benefit derived from the provision of C voltage reduction and from the stabilization of receiver performance at low battery voltages, some remarkable increases in the useful life of the B batteries are to be obtained. In some instances the useful life of the B batteries may be more than doubled, while a conservative average for such increases is of the order of 50 per cent, representing a saving in the operating cost of the receiver which is repeated each time the consumer buys a new set of batteries.

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THE DEVELOPMENT AND APPLICATION OF MARINE RADIO DIRECTION FINDING EQUIPMENT BY THE UNITED STATES COAST GUARD*

By

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Summary—*The various problems dealt with in the evolution, development, and practical application of radio direction finding equipment by the United States Coast Guard are discussed. The equipment in use at the present time is described. Difficulties encountered with marine equipment due to the electrophysical properties of vessels are explained. The methods employed in overcoming these difficulties are described.*

Some notes on deviation are presented along with curves and photographs showing the results obtained on various types of vessels.

Deviation as a function of frequency is discussed in brief.

The material presented is based on data obtained from more than one hundred direction finder installations on vessels of various types and sizes ranging from 75-foot motor boats to cutters of 3000 tons displacement.

A brief description of aircraft equipment and an account of the results obtained therewith is included.

The object of this paper is to present a résumé of the results obtained with modern equipment under the actual conditions encountered in service use, accompanied by such notes and comments by the author as are deemed of interest to those concerned with the development, improvement, and application of marine and aircraft radio direction finders. The fundamental principles underlying the art of direction finding by means of radio are not recounted as it is realized the reader has available numerous current publications on the subject.

This paper is published by permission of the Commandant, United States Coast Guard. The data presented have been obtained during the development, improvement, and installation of equipment used on vessels and aircraft of the Coast Guard. All of the equipment described was manufactured in accordance with specifications prepared and issued by that Department. All photographs are the property of the Coast Guard and are not to be reproduced without permission.

THE manifold duties of the United States Coast Guard, particularly those concerning the preservation of life and property on the seas and the Great Lakes, render mandatory the use of efficient radio direction finding equipment. In order that assistance to distressed craft might be accomplished with a minimum loss of time regardless of weather conditions, all seagoing vessels of the Coast Guard are equipped with radio direction finders of the most modern type. The useful

* Decimal classification: R152×R325.31. Original manuscript received by the Institute, September 3, 1931. Presented before New York Meeting, October 7, 1931.

application of this equipment, however, is not by any means limited to distress relief work. The marine radio direction finder is as consistently employed by the present-day Coast Guard navigator as the compass, pelorus, and sextant. In conducting a constant patrol along the shores of the United States, the annual international ice patrol, and cruising in Alaskan waters, it is frequently necessary for Coast Guard vessels to make physical contact with each other during adverse weather conditions when, due to poor visibility, the reliable range of most navigational devices is greatly limited. Under these circumstances it would at times be impossible to accomplish contacts without the aid of accurate radio direction finding equipment.

I. Evolution of Equipment

The development of equipment capable of satisfactory operation on the various types of Coast Guard vessels which range from 210 to 3000 tons displacement involved the consideration of numerous electrical and mechanical details embodied in the structural make-up of the vessels. Space limitation was of course a prime consideration. Since the radio direction finder is a navigational instrument, its logical location is on or near a vessel's bridge. Those familiar with the structural arrangements usually employed on the bridges of small- and medium-sized vessels will readily understand the necessity for imposing dimensional restrictions on equipment designed for bridge or pilot house use. The problem is electrically complicated due to the proximity of numerous vertical conductors in the form of stays, guys, mast, stacks, railing, stanchions, shrouds, etc., which are usually located in the vicinity of the bridge and radically affect the operation of the radio direction finder. The rotable loop-type direction finder was chosen as the most practicable arrangement for marine use. Early attempts were not entirely satisfactory due to inherent faults of the equipment, principally its susceptibility to local distortional influences and the tendency of the circuit consisting of the loop antenna, receiver, and ground to function as a simple vertical antenna. Earlier models consisted of a rather large frame loop connected to a radio receiver usually of the untuned radio-frequency type. The loop, being quite large and unshielded, was highly responsive to local induction fields, particularly those resulting from lengthy vertical objects of a metallic nature such as standing rigging and masts with their associated equipment, the electric cables, voice tubes, ladders, etc., the effects of which are apparent in the form of displacement errors and obscure minima.

The currents induced in a loop-type radio direction finder by local metallic objects, which form closed loops, produce spurious fields

which when added vectorially to the desired signal, produce a resultant field which may differ materially in direction from the direction of arrival of the incoming wave. This difference between the real and apparent direction is called the displacement error or the deviation. Currents induced in the direction finder by local objects which act as vertical antennas are, in general, 90 degrees out of phase with the current induced by the received radio signal. As no value of signal currents can be obtained by rotating the direction finder loop to cancel out these spurious currents, obscure minima result in opposite areas, particularly in the beam sectors, as the plane of the direction finder loop is parallel to the vessel's fore and aft axis and a position of maximum flux linkage with mast and stack fields is attained. Original efforts to overcome or neutralize the above effects were limited to breaking up the continuity of near-by metallic objects in so far as practicable by the insertion of insulators. As it is not practicable, however, to destroy the electrical continuity of such objects as electric cables and lightning conductors located on the masts, fields set up by these objects still remained and were apparent to the observer in the form of obscure minima.

The displacement error, or deviation[1] as it is more popularly termed, is seldom of sufficient magnitude to merit special electrical attention. The major problem, and by far the most difficult one to solve with marine radio direction finder installations, is to obtain pure, well-defined minima on signals approaching from any angle relative to the vessel's bow, particularly in the beam quadrants. An analysis of the results obtained with more than one hundred installations show that displacement error and minima classification according to type and size of vessels is not possible. For instance, two cutters or two destroyers of identical construction and size frequently show strikingly dissimilar direction finder operating characteristics. This inconsistency tends to render more difficult the design of certain electrical parts of the direction finder, particularly the arrangement for balancing out residual signal[2] due to field effects from local objects which act as vertical antennas. This balancing is accomplished by introducing, magnetically or capacitively, currents of definite phase value into the loop circuit which "balance out" the undesired residual signal. The evolution of the present-day loop-balancing system employed in Coast Guard direction finders is shown by Fig. 1 in which (a), (b), and (c) show progressive arrangements, the first of which consisted merely of a variable condenser from grid to ground which compensated for the

[1] In the case of a vessel, the difference between the radio and visual bearing is measured relative to the bow.

[2] That signal which sometimes exists in the null area which should be well defined and void of all signal.

comparatively high capacity to ground on the filament side.[3] This arrangement was employed with the early type, large, unshielded, open-frame loops which were notoriously unsymmetrical and highly respon-

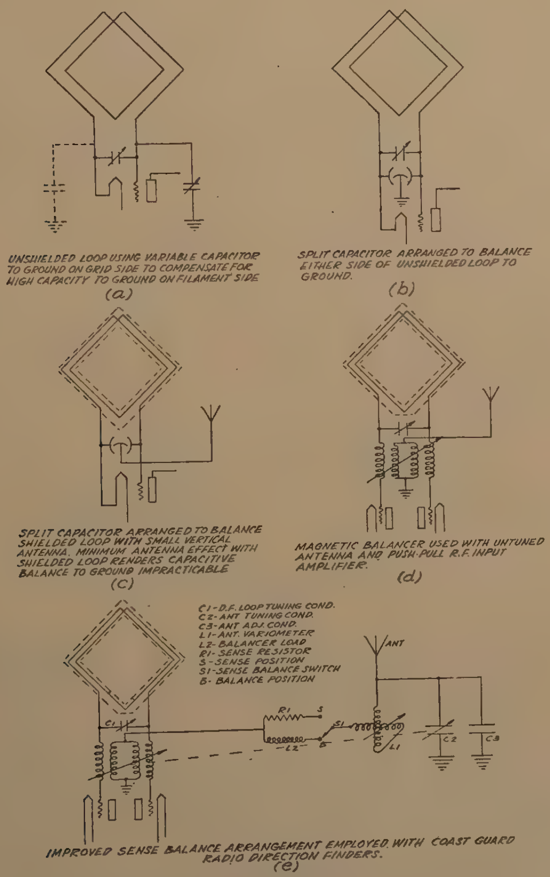


Fig. 1—Evolution of sense-balancer circuit arrangements.

sive to local static fields. Fig. 1 (b) shows the next arrangement devised which consisted of a split variable capacitor to ground. This scheme was a great improvement over the former method but its application was found to be limited due to tuning effect introduced across the direction finder loop. Also the size of the condenser required to balance out exceptionally strong residual signals proved too large to be

[3] Ballantine in 1917–1920 applied this device to U. S. Navy loop-type direction finders as a "capacity compensator" for the purpose of reducing "antenna effect."

mechanically practicable. The balancer problem was greatly simplified by the advent of the electrostatically shielded loop and the use of push-pull r-f amplification shown by Fig. 1 (c), (d), and (e). It was found that shielding the loop rendered it far less susceptible to local distortional influences and tended to preserve its electrical symmetry. The capacitive method of balancing shown by Fig. 1 (c) was at first employed with the shielded loop but, due to the tendency of such arrangement to tune the loop circuit, it was eventually superseded by the magnetic type, Fig. 1 (d), which consisted of a split winding balancer arranged for imparting signal energy to either side of the loop circuit from

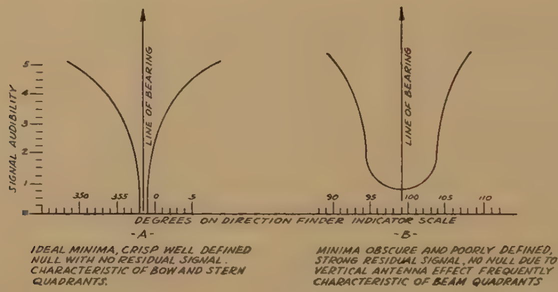


Fig. 2—Residual signal.

a simple vertical antenna. The rotor inductors of this balancer were wound so that their fields assisted each other. An electrostatic shield was placed between the rotor and stator windings. This arrangement did not tune the loop unless a resonant condition was attained between the loop and antenna circuits due to excessive antenna pick-up resulting from overcoupling. To prevent this resonance occurring under the various conditions encountered on board ship, while at the same time obtaining satisfactory balancer effect over the entire frequency range covered by the receiver, constituted quite a problem but was finally accomplished by tuning the balancer circuit as shown by Fig. 1 (e) which shows the arrangement employed with present-day equipment. The two windings of the balancer rotor are in field opposition to each other thus eliminating any tendency to tune the loop circuit inductively. The loop tuning condenser, C, operates in tandem with the antenna tuning condenser, C_2 and variometer, L, by means of a common manual control. This arrangement permits the same comparative ratio of the loop and balancer antenna values over the entire frequency range while at the same time preventing a state of resonance between the two. The sense-balance switch, S, is normally in position B, which connects

the antenna circuit to the balancer variometer through the loading inductance L_2. When the switch is placed in position S, contact is completed through the resistor R, unbalance effect being obtained to distort sufficiently the normal figure-of-eight loop pattern in obtaining directional sense indication. As Coast Guard radio direction finders oper-

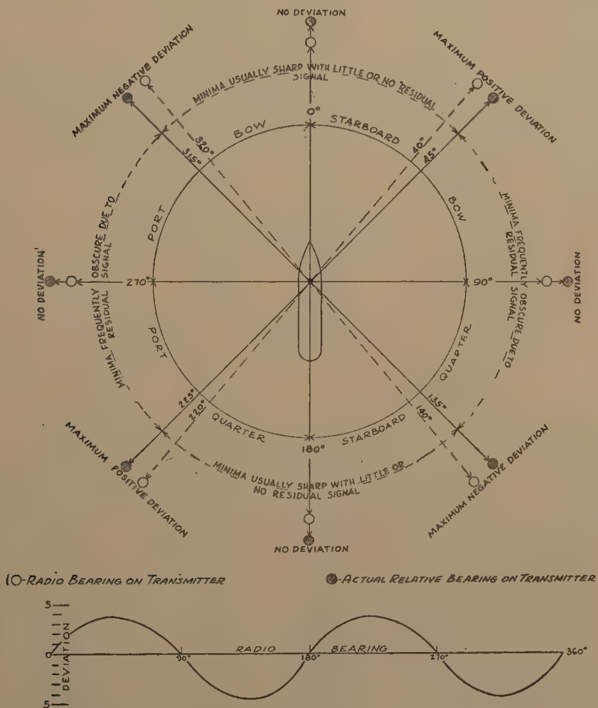


Fig. 3—Normal quadrantal error and minima characteristic.

ate on the principle that a loop antenna will indicate the *line of bearing* of an observed radio signal by diametrically opposite minima it is necessary to provide a means for indicating the *direction* from which the signal is approaching. This is accomplished by imparting currents from the balancer antenna to the direction finder loop which are phased so as to assist the loop currents when the plane of the loop lies in the direction of signal approach but opposes the loop currents when the plane of the loop is reversed 180 degrees. The effect apparent to the observer is two opposite but audibly unequal areas of maximum signal intensity. The resistance R is of such value that the required phase change is accomplished when the switch is placed on the S contact. The values L_1, L_2, C_2, and C_3 are such that the currents in the balancer circuit are

capacitively, inductively, and resistively[4] phased to give satisfactory balance effect without permitting a state of resonance between the loop and balancer circuits. Early experience showed that under certain conditions such resonant or near-resonant relationship might create false minima by the balancer adjustment shifting the apparent position of

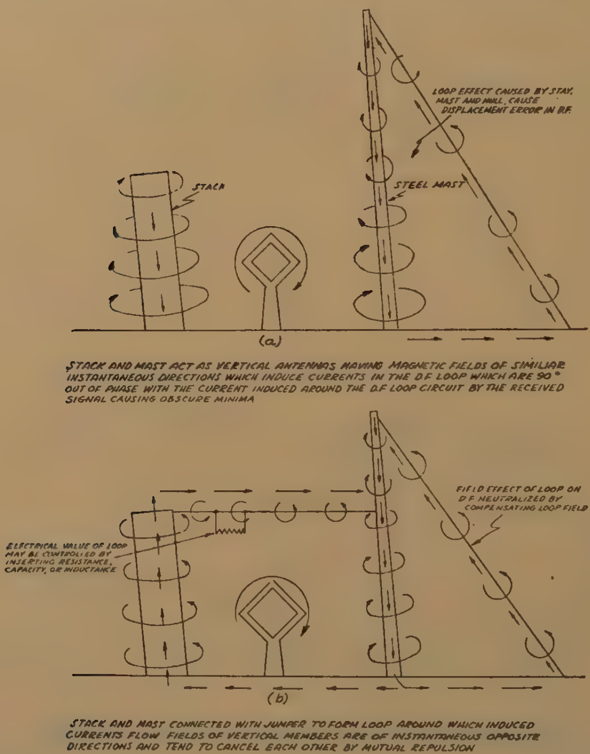


Fig. 4—Reduction of quadrantal error and improvement of null quality by means of compensating loop.

the null regardless of the relative angle of the plane of the loop with regards to the direction of signal approach. This effect is not possible with the perfected arrangement shown in Fig. 1 (e). Without a means of balancing out spurious loop currents it would not be possible to obtain pure, well-defined minima under the adverse conditions frequently met with in marine radio direction finder installations. Residual signal as apparent to the direction finder operator is illustrated by Fig. 2.

Strong residual signal and excessive deviation (displacement error) are not directly associated although both are frequently present to-

[4] Inherent circuit resistance.

gether, especially in the beam quadrants. Fig. 3 illustrates typical quadrantal displacement and null quality experienced at low and inter-

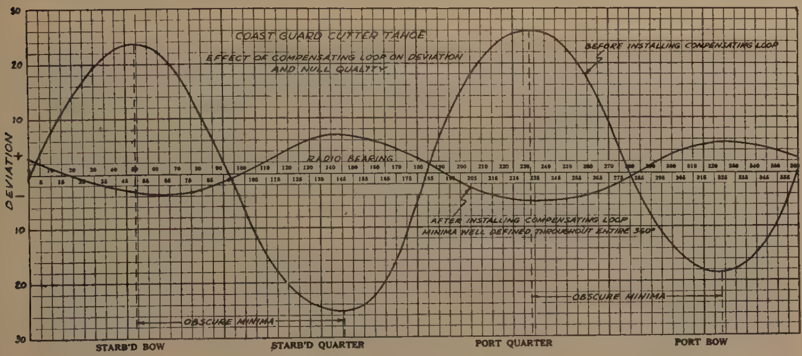


Fig. 5.

mediate frequencies with marine radio direction finders. In the case illustrated it is assumed that no balancing system is employed. The ob-

Fig. 6—U. S. Coast Guard Cutter Tahoe.

scure null sectors shown would be corrected in actual practice by employing an efficient balancer.

II. Reducing Excessive Deviation

Excessive deviation can be effectively reduced by employing a compensating or current loop as explained in Fig. 4. It has been found that the use of an arrangement of this sort not only reduces deviation but

also improves the null quality in obscure sectors as further illustrated in Fig. 5. Under normal conditions a compensating loop is not required and is not employed unless the deviation is excessive, i.e., more than

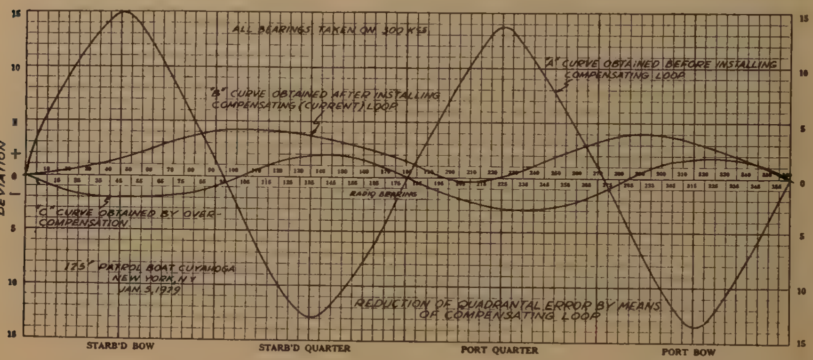


Fig. 7.

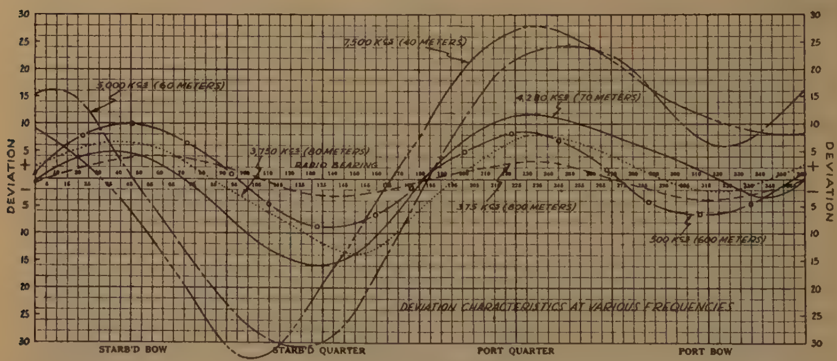


Fig. 8.

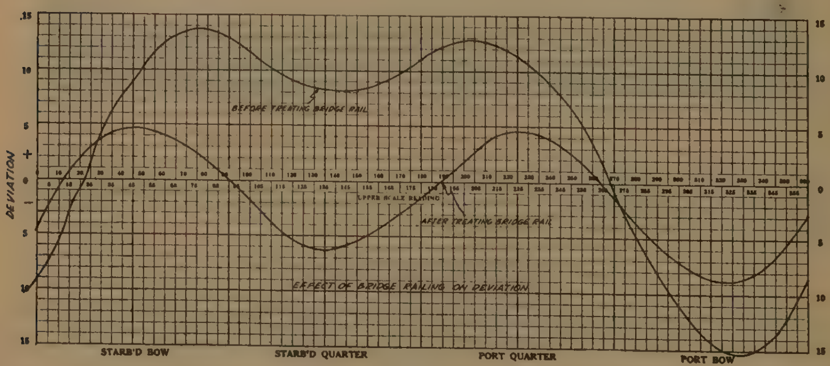


Fig. 9.

20 degrees, which is beyond the practical range of mechanical compensation (see Fig. 25). If the deviation is normal but obscure sectors exist, such condition can be corrected by employing a balance antenna

Fig. 10—U. S. Coast Guard Cutter Yamacraw.

of sufficient size in conjunction with the proper values required to induce currents of sufficient magnitude to balance out the residual signal.

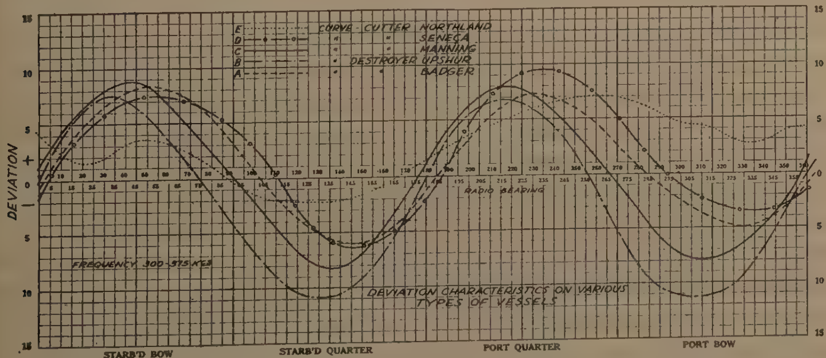


Fig. 11.

The electrophysical dimensions of the compensating loop are extremely important and must be carefully determined at the time of installation. If a critical value near that of the direction finder frequency is

attained the compensating loop will become resonant with the direction finder and enormous displacement error will result. If a loop of too restricted dimensions is employed an overcompensated condition such as that illustrated by curve *C* of Fig. 7 will result. Where physical limitations prevent installing a compensating loop of sufficient size, it may be adjusted to the required electrical value by inserting inductance, capacity, or resistance at some convenient point as indicated by Fig. 4 (b).

Fig. 12—U. S. Coast Guard Destroyer Abel P. Upshur.

The cutter Tahoe shown in Fig. 6 is a typical example of the effectiveness of a compensating loop in reducing excessive quadrantal error and the elimination of residual signal in the beam quadrants. In this case the direction finder loop is installed on the upper bridge in the midst of numerous metallic objects. The exceptionally tall mast with its associated rigging and the electric wiring, voice tubing, etc., set up such a strong field that the deviation amounted to more than twenty degrees attended by obscure minima in the beam quadrants. The installation of a compensating loop, which merely consisted of connecting the metallic objects on the mast to the forward bridge railing by means of the bonding jumper, shown in Fig. 6, extending from a point about ten feet below the lookout to the bridge railing reduced the average error from 23 degrees to 5 degrees and eliminated most of the residual signal in the beam quadrants. This jumper completed a triangular-shaped loop formed by the mast conductors, the metallic superstruc-

ture, and the jumper itself. The plane of the compensating loop thus formed is fore and aft and encloses the direction finder loop. The results obtained are comparatively shown by the curves in Fig. 5. The curve showing deviation after installing the compensating loop indicates a somewhat overcompensated condition which could be corrected by tuning the compensating loop or by increasing its physical dimensions.

Fig. 13—U. S. Coast Guard Cutter Northland.

This curve, however, is substantially flat and well within the corrective scope of the mechanical deviation corrector which comprises part of all standard Coast Guard direction finders. (See Fig. 25.)

In rare instances the fields set up by horizontal members of metallic railing and awning support systems installed in the vicinity of the direction finder have been found to be the cause of peculiar deviations. An illustration of a case of this nature is shown by Fig. 9 in which curve "*A*" indicates the abnormal deviation before the electrical continuity of the bridge railing and awning support system on the cutter Yamacraw (Fig. 10) was destroyed. Curve "*B*" shows the satisfactory results obtained after breaking up the horizontal members of the awning support and railing arrangement. Similar cases are rarely encountered as the physical dimensions of local metallic objects located in a horizontal plane are usually so small that their natural frequency is well below that at which the direction finder is operated.

III. Deviation Characteristics on Various Types of Vessels

The contour and magnitude of the deviation is more or less different for each individual vessel and cannot be classified according to any particular class or type as may be seen by comparing the curves in Fig. 11 showing the deviation noted on the destroyers Abel P. Upshur and George E. Badger, which vessels are identical in type and physical dimensions. One of these destroyers, the Abel P. Upshur is shown in Fig. 12. The location of the direction finders on these two vessels are exactly the same and the rigging and mast arrangements are identical, yet the difference in deviation is considerable. It is further interesting to note that the null quality in the beam quadrants of the Abel P. Upshur was excellent while in the same sectors on the George E. Badger the null showed considerable residual signal which to balance out required an antenna considerably larger than the one employed on the Upshur. The deviation curves of these two destroyers are substantially symmetrical and purely quadrantal as the direction finders in each case are located near the center line. The irregular curve obtained on the cutter Northland (Figs. 11 and 13) is due to the extensive rigging arrangement on the foremast and the peculiar contour of the vessels hull and superstructure. The Northland was constructed for crusing in Alaskan waters and is particularly designed for work in the ice fields. She has a length of 216 feet over-all, her displacement is 2050 tons with a maxi-

Fig. 14—U. S. Coast Guard Cutter Manning.

mum draft of 15 feet. The original direction finder installed on this vessel employed an open-frame unshielded loop which proved unsatisfactory. Obscure minima prevailed throughout the beam sectors in spite of any balance antenna or compensating loop arrangement physically possible to employ. The present installation consists of a standard CGR-17-B direction finder equipped with a shielded loop which operates satisfactorily. By means of the shielded loop combined with a com-

Fig. 15—U. S. Coast Guard Cutter Seneca.

pensating loop and balance antenna of carefully determined electrophysical value, well-defined minima are obtained in the sectors which were obscure when using an unshielded loop. This is only one of several instances encountered in marine installations where satisfactory results could not be obtained with an unshielded loop. No case, however, has ever been found where a shielded loop used in conjunction with the proper compensating loop and balance antenna has failed to give the required results. As a contrast to the results obtained on the Northland, the cutter Manning, shown in Fig. 14, is mentioned as an ideal installation from a viewpoint of favorable local conditions. This vessel has a length of 205 feet, a beam of 32 feet and a displacement of 1155 tons. The deviation obtained on this vessel is shown in Fig. 11. The symmetrical contour is due to the direction finder's being located on the center line. The original installation made in 1924 employed an open-frame unshielded loop which worked very well due to the exceptionally favorable environment. Sharp, well-defined nulls with no residual signal were obtained throughout the entire azimuth. The directional sense in-

dication was excellent. The Seneca, (Fig. 15), a cutter of approximately the same size as the Manning, but having only one mast, although possessing good direction finder operating qualities as to deviation, null quality, and directional sense indication (Fig. 11) was not as satisfactory as the Manning.

IV. Deviation Characteristics at Various Frequencies

Radio direction finders developed by the Coast Guard for navigating and contacting purposes are designed to operate over a frequency

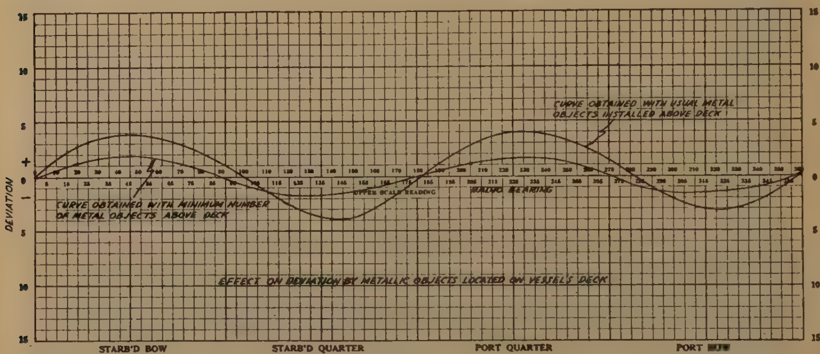


Fig. 16.

range from 275 to 550 kilocycles. This range takes in the distress, radio direction finding, and radio beacon frequencies which are all that are required for regular service operating purposes. The possibilities, however, of equipment intended for special purposes to operate on higher frequencies have been investigated. Some of the results obtained in this field are described particularly to show the contrasting deviations noted at various frequencies. The six curves shown in Fig. 8 indicate the nature of the deviation noted at various frequencies ranging from 375 to 7500 kilocycles. It will be readily noted that the deviation at the lower frequencies is quadrantal in form while at the intermediate frequencies it becomes increasingly semicircular in form and finally, at 7500 kilocycles, it becomes definitely semicircular. This tendency, it is believed, holds good on all types of vessels, the amplitudes of course being proportional to the vessel's size and the nature of metallic objects located above the decks and near the direction finder. The results shown by Fig. 8 were obtained with special equipment installed on a 75-foot patrol boat equipped for the purpose of conducting radio direction finder tests over a wide range of frequencies.

That the usual quadrantal error experienced at the intermediate

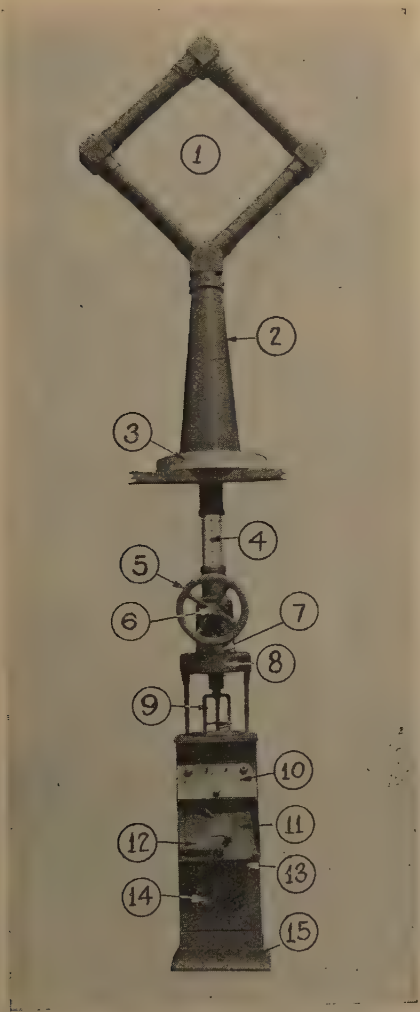


Fig. 17—U. S. Coast Guard radio direction finder, Type-A, Model CGR-17-C.
1. Loop.
2. Loop supporting pedestal.
3. Pedestal mounting blocks.
4. Extension tube.
5. Hand wheel.
6. Locking clamp.
7. Drive housing.
8. Mechanical compensator housing.
9. Indicator.
10. Control panel.
11. Receiver door.
12. Clip for telephones.
13. Receiver cabinet.
14. Lower compartment door.
15. Receiver base.

Fig. 18—U. S. Coast Guard radio direction finder, Type-A, Model CGR-17-C, lower assembly.

1. Extension tube.
2. Hand wheel.
3. Lock.
4. Drive and compensator housing.
5. Indicator.
6. Receiver panel.
7. Clip for telephones.
8. Lower compartment door.
9. Receiver base.

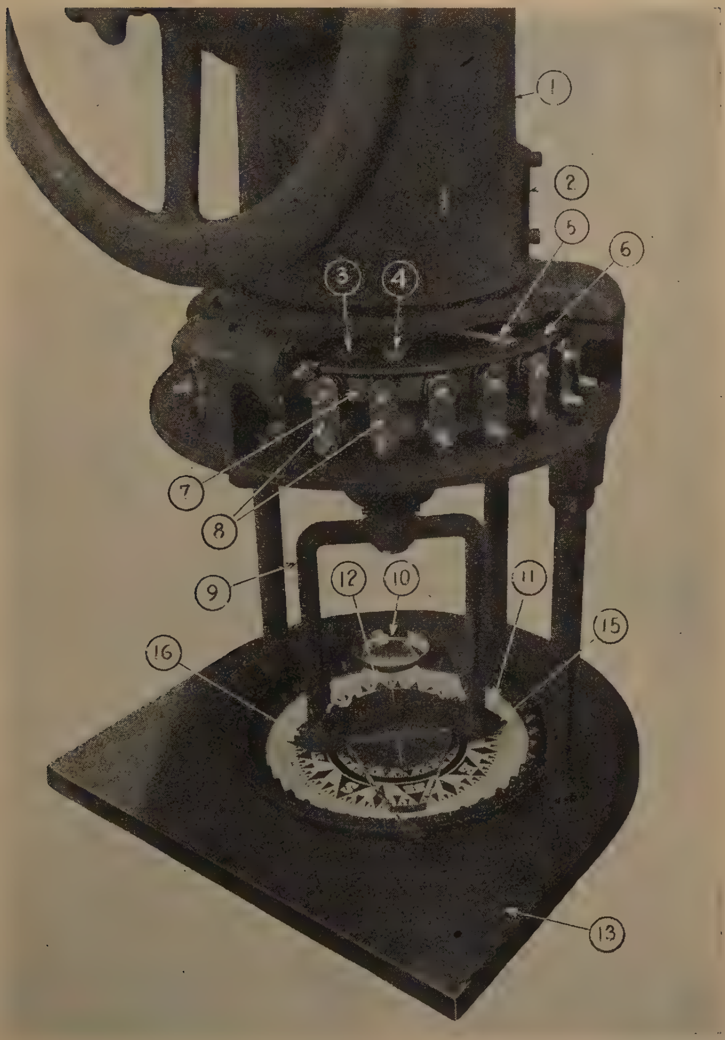


Fig. 19—U. S. Coast Guard radio direction finder, Type-A, Model CGR-17-C. mechanical compensator and indicator arrangement.

1. Drive housing.
2. Collector rings plate.
3. Tension spring.
4. Loop pointer.
5. Add 40 degrees loop pointer.
6. Loop, calibrating, or upper scale.
7. Flexible band track.
8. Track adjusting clamps.
9. Indicator yoke.
10. Reading glass.
11. Dumb compass.
12. Sight vane glass.
13. Receiver cabinet top.
15. Maximum "red" pointer.
16. Maximum "white" pointer.

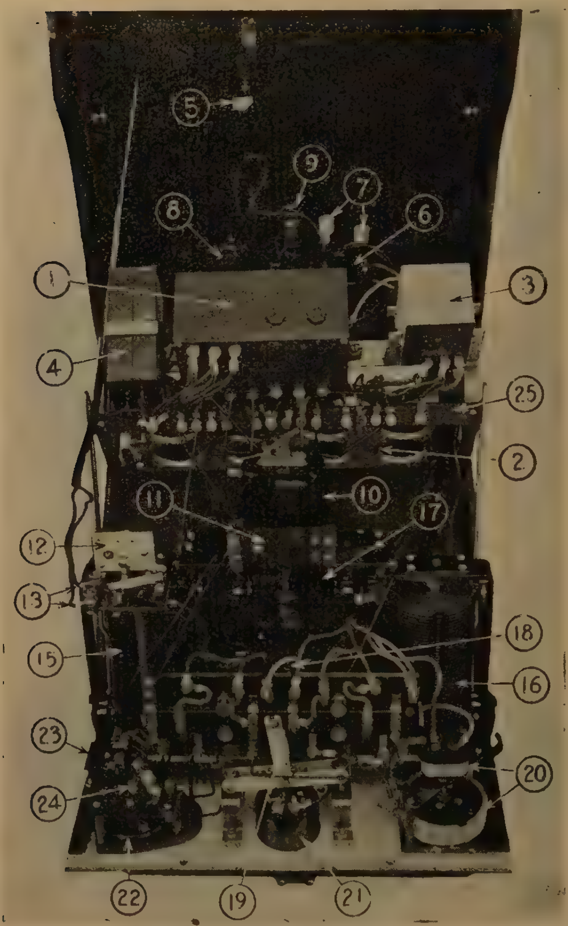


Fig. 20—U. S. Coast Guard radio direction finder, Type-A, Model CGR-17 C, receiver assembly.

1. Coupling and oscillator unit.
2. Catacomb.
3. CW oscillator unit.
4. Loop by-pass condenser.
5. Charging resistor.
6. Relay.
7. Sockets for spare lamps.
8. Lightning arrester.
9. Repeater cable tube.
10. Sense-balance variometer.
11. Antenna trimming condenser.
12. Loop trimming condenser.
13. Loop leads and terminals.
14. Loop bias resistors.
15. Loop tuning condenser.
16. Oscillator tuning condenser.
17. Antenna tuning condenser.
18. Radio-frequency tuning condenser.
19. Master switch.
20. Rheostat and cw switch.
21. Filament-plate voltmeter.
22. Balancer.
23. Balancer load coil.
24. Sense resistor.

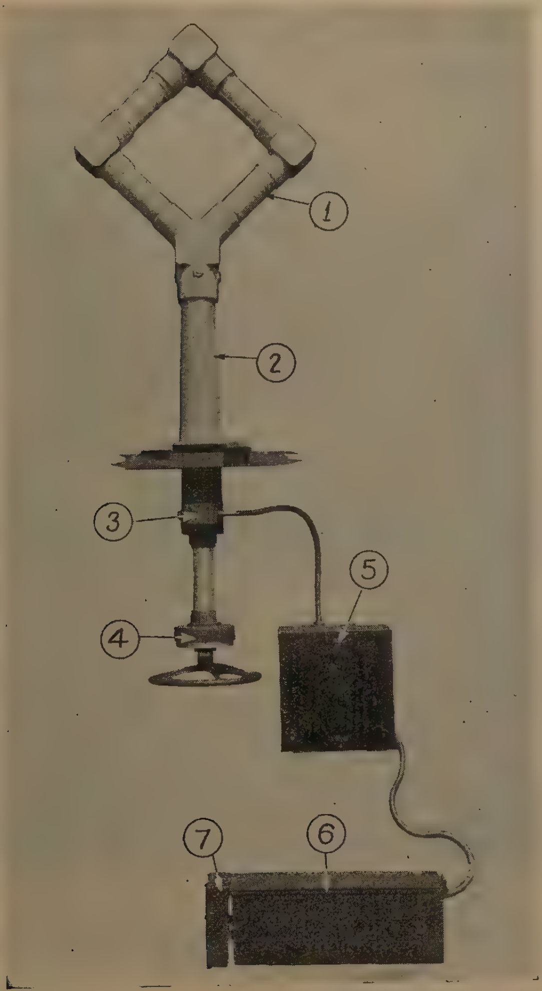


Fig. 21—U. S. Coast Guard radio direction finder, Type-B, Model CGR-18, complete assembly.

1. Loop.
2. Loop pedestal.
3. Commutator.
4. Compensator and drive.
5. Receiver.
6. Battery box.
7. Charging unit.

and lower frequencies with marine radio direction finders is primarily due to the electrophysical characteristic of the vessel's hull but is augmented by the fields resulting from stays, mast, stack, etc., is indicated by the results graphically illustrated in Fig. 16. The curves shown were obtained at a frequency of 2705 kilocycles. Curve "*B*" indicates the deviation noted with the usual metallic objects found abovedeck situ-

Fig. 22—U. S. Coast Guard radio direction finder, Type-B, Model CGR-18, receiver.

ated in their regular places and partially treated to minimize field effects. Curve "*A*" shows the result of removing these objects and indicates the deviation resulting from the collective distortional effects of metallic objects and equipment located within the vessel's hull, which in this case was constructed of wood. If the hull were of metal the quadrantal characteristic at this frequency would no doubt be more pronounced. That quadrantal error is primarily due to the refractive influence of a vessel's hull rather than the masts, stack, rigging, etc., is substantiated by the fact that errors of a similar type can be produced

on land by placing the direction finder on top of a building or on a hill. The latter effect is mentioned in an article by R. L. Smith-Rose.[5] Mention is further made in the above article that direction finder errors

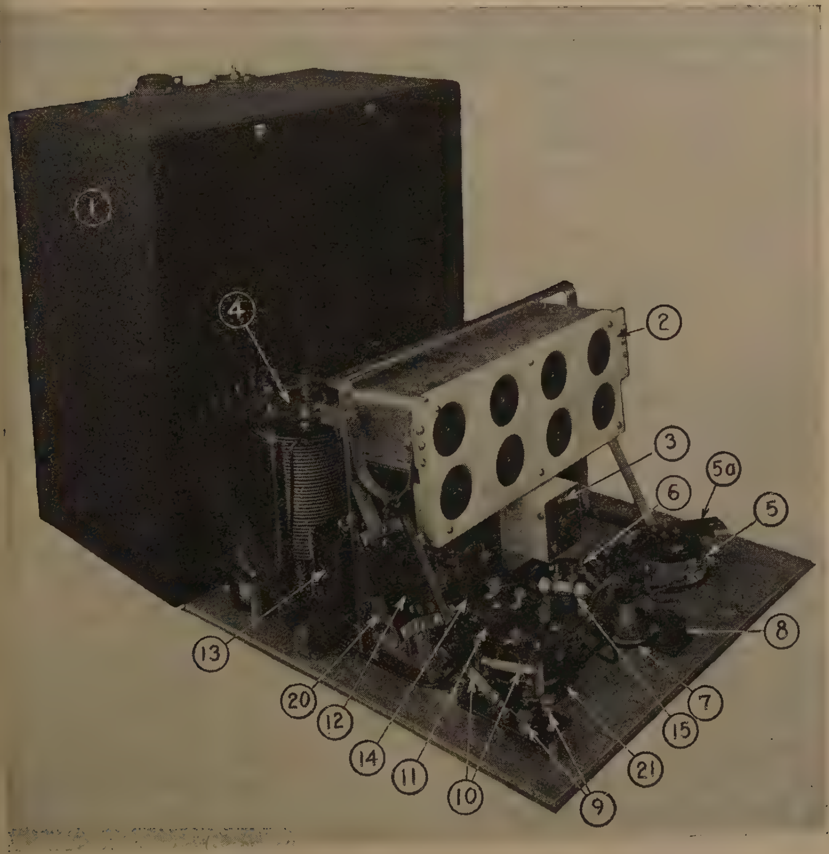


Fig. 23—U. S. Coast Guard radio direction finder, Type-B, Model CGR-18, receiver assembly.

1. Cabinet.
2. Catacomb.
3. CW oscillator unit.
4. Loop tuning condensers.
5. Filament rheostat and antenna grounding switch.
6. Voltmeter.
7. Panel light socket.
8. Illumination switch.
9. Loop terminals.
10. Bias resistors.
11. Sense-loading inductance.
12. Tuning trimmer condenser.
13. Fixed condenser.
14. High-frequency antenna coupler.
15. Sense resistor
20. Coupling unit
21. Balancer and sense switch.

were found to be independent of the wavelength. The results described in this paper, however, are not in accord with the Smith-Rose claim regarding deviation as not being a function of the frequency. That fre-

[5] R. L. Smith-Rose, Proc. I.R.E., March, 1929.

quency does affect deviation has been established as a very definite and highly important fact and accounts for the standard practice in the Coast Guard of determining the deviation, even on frequencies of the lower order of 300 to 500 kilocycles which are used for regular navigational radio direction finder purposes by calibrating at each end and the middle of that band in all cases where the deviation is excessive or

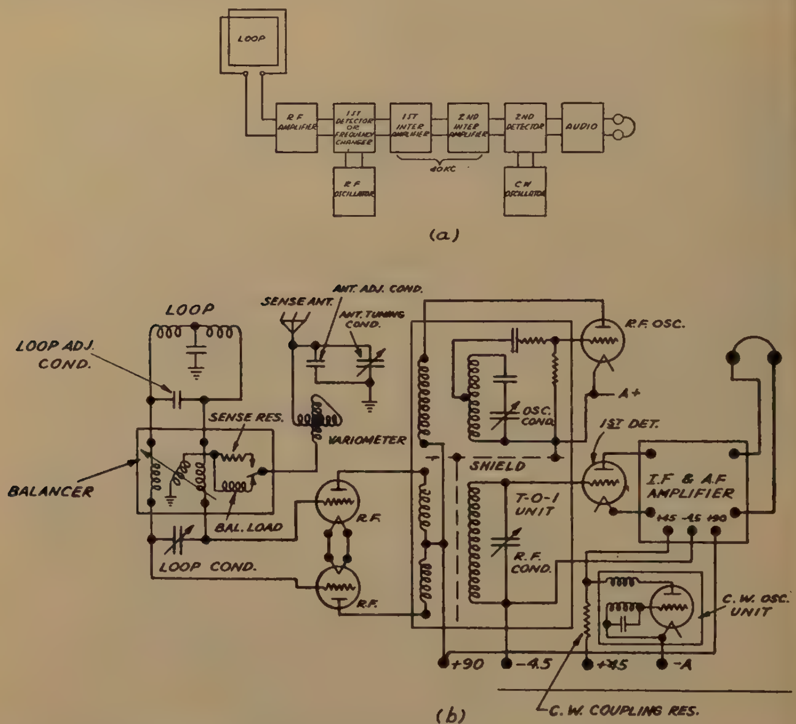
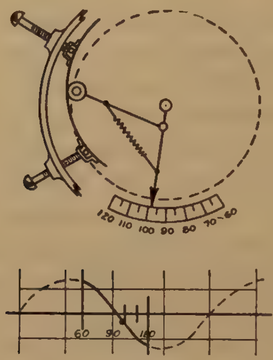


Fig. 24—U. S. Coast Guard radio direction finder schematic circuit arrangement

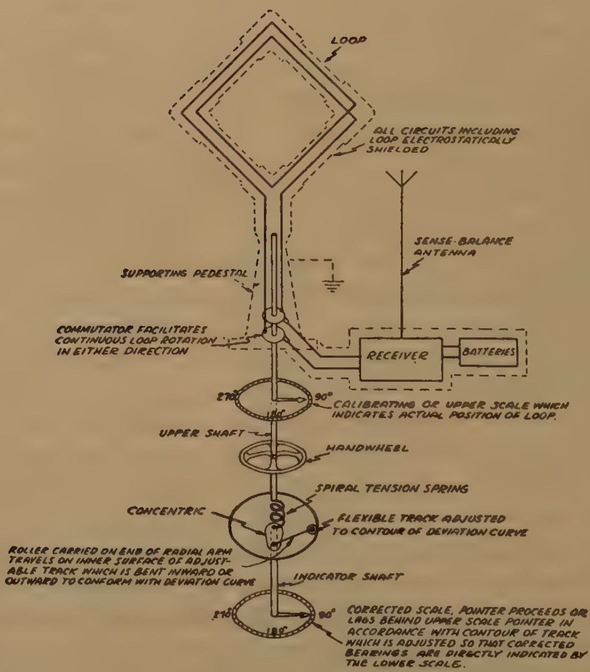
erratic. If the deviation at 300 kilocycles is found to be normal and does not exceed approximately ten degrees, experience has shown that a calibration made on this frequency is, as a general rule, sufficiently accurate for 500 kilocycles. Cases are infrequently experienced where the displacement error at 300 kilocycles differs radically from the error at 500 kilocycles. Critical deviation may be likened to a steel spring several feet long held at each end between a person's hands. If the angle and relative pressure at which the ends are held are varied the spring will readily assume various patterns. Such is the case with excessive deviations or those of erratic contour, which necessitate calibrating the direction finder on each frequency to be used for taking bearings.

V. Standard Coast Guard Marine Equipment

Standard direction finders installed on Coast Guard vessels are of two types, designated as A and B respectively. Each type includes several models. Type A equipment is designed for use on vessels of the larger type such as cutters and destroyers. (Fig. 17.) Type B equipment is designed to meet the requirements of smaller craft such as the 100-foot, 125-foot, and 165-foot vessels used for patrol work. (Fig. 21.) Both types employ the same circuit arrangements but differ somewhat mechanically and in size in order to meet certain physical requirements. A Type A, Model CGR-17-C, direction finder, completely assembled is shown by Fig. 17. Figs. 18, 19, and 20 are detailed views of the various parts comprising the equipment. Fig. 21 shows the complete assembly of a Type B, Model CGR-18, direction finder. Figs. 22 and 23 are close-up views of the Model CGR-18 receiver. The circuit employed by both types is the double detection or so-called superheterodyne arrangement schematically shown in Fig. 24. The loop leads feed into a pair of vacuum tubes arranged as push-pull r-f amplifiers. This system affords good loop balance and tends to preserve its symmetry by eliminating a high capacity side to ground which would be the case if the loop fed directly into a single stage of r-f amplification or into the first detector. The remainder of the circuit consists of a first detector, r-f oscillator, two stages of 40-kc intermediate-frequency amplification, cw oscillator, second detector, and one stage of audio-frequency amplification. In all of the later models, tuning is accomplished by means of one manual control which also actuates the balance antenna tuning. The frequency range covered is from 275 to 550 kilocycles, continuously variable, and provision is made for reception of cw or icw signals. The Type B, Model CGR-18, in addition to functioning as a radio direction finder on the above frequency band is arranged to operate as a non-directional radio receiver on the Coast Guard high-frequency band of 2300 to 2700 kilocycles. Transfer from low-frequency direction finder to high-frequency receiver is accomplished by means of the master switch, the control of which is shown on the lower center of the panel in Fig. 22. Correction for the deviations noted at time of calibration is accomplished by means of a mechanical device which comprises part of all standard models and is so arranged that the indicator sight vanes are caused to precede or lag behind the plane of the direction finder loop in the amount necessary to compensate for the displacement error throughout the entire azimuth. The action of this corrector, or mechanical compensator, as it is more appropriately termed, is entirely automatic so that bearings read directly from the indicator are corrected for quadrantal or semicircular deviation and are not referred to



(a)



(b)

Fig. 25—Operating principle of mechanical deviation corrector employed with U. S. Coast Guard radio direction finders.

a chart or calibration curve. This method of mechanical compensation must not be confused with electrical compensation described in a preceding paragraph. Fig. 25 illustrates the principle of mechanical compensation employed with Coast Guard radio direction finders. Type A equipment is so arranged that the dumb compass, shown as part of the indicator in Fig. 19, may be replaced by a gyro-repeater on vessels equipped with a gyro-compass in order that bearings taken with the

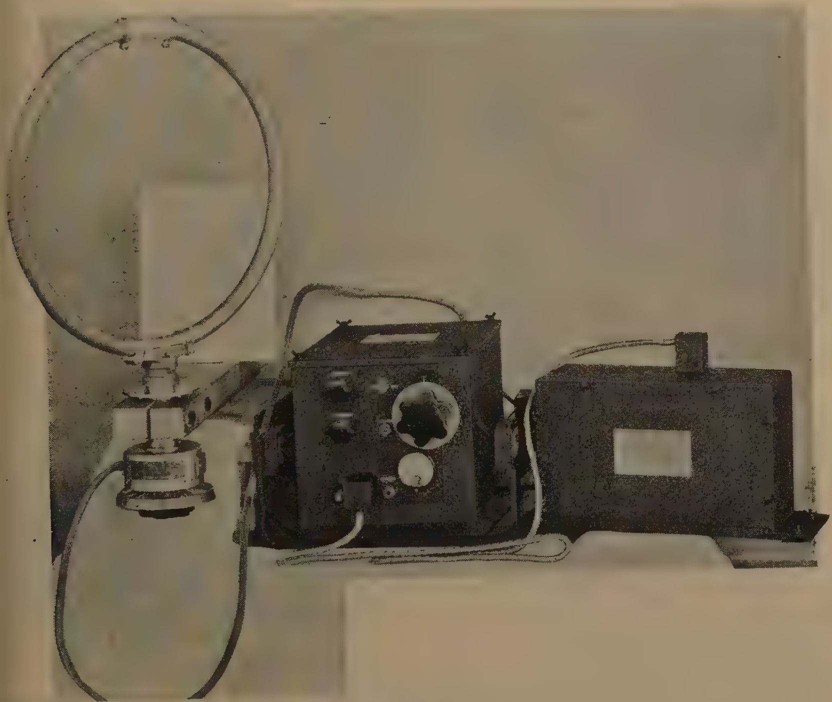


Fig. 26—U. S. Coast Guard aircraft radio direction finder, Model CGR-19-B.

direction finder are *true* bearings as read directly from the indicator. This arrangement eliminates the possibility of observational errors and those due to yawing of the vessel mentioned by R. L. Smith-Rose.[6] All first-class cruising cutters and destroyers of the Coast Guard are gyro-compass equipped. The smaller vessels which do not have gyro-compasses are provided with the Type CGR-18 direction finder which has the indicator so arranged that the true, relative, or magnetic significance of the observed radio bearing is directly indicated. This is accomplished by having the indicator scale movable independently of the loop shaft. A center line or "lubber's line" mark, rigidly attached to the

[6] *See* footnote 5.

loop drive, is located immediately above the indicator scale. To obtain a bearing relative to the vessel's bow the indicator scale zero is set opposite the lubber line mark. The direction finder now functions as a radio pelorus. To obtain a true or magnetic bearing the indicator scale is simply set with the true or magnetic heading opposite the lubber line mark. It is essential of course that the vessel be exactly on her course when using the direction finder in this manner. A "mark" signal is al-

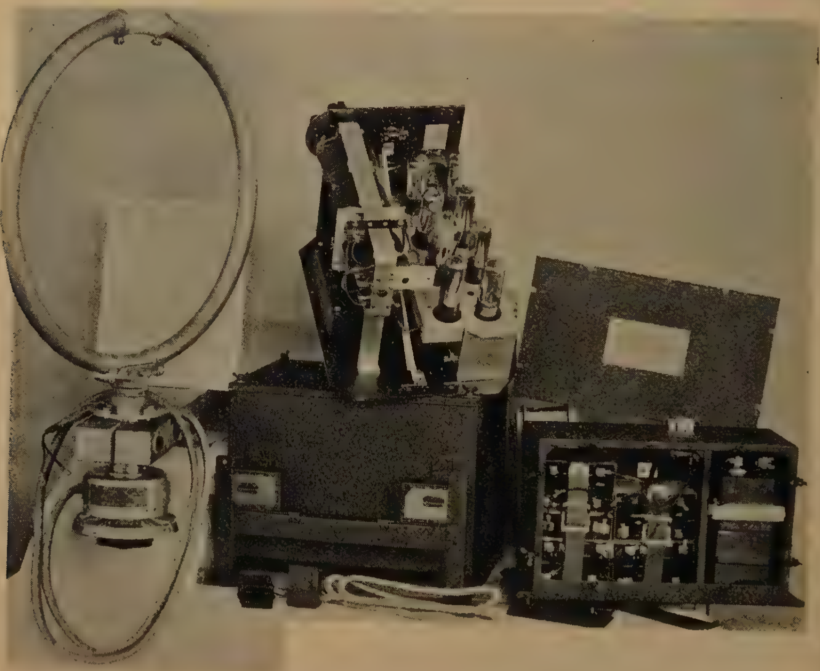


Fig. 27—U. S. Coast Guard radio direction finder for aircraft, Model CGR-19-B.

ways given by the radio direction finder operator at the instant the bearing is observed in order that the ship's head may be carefully noted and allowance made for yawing. Reduction of direct pick-up is accomplished with Coast Guard direction finders by including the loop, loop leads, receiver, and all accessories within a common electrostatic shield, the continuity of which is broken at the apex of the loop (Figs. 17 and 21) by means of insulating fittings. Shielding the loop reduces its pick-up somewhat but this sacrifice in sensitivity is well worth the advantage gained by reduced response to local fields.

VI. Aircraft Equipment

A rotable loop-type direction finder developed for use on Coast Guard aircraft employs the same circuit arrangement as the standard

marine equipment except that one stage of screen-grid r-f amplification is employed in the input circuit instead of two balanced stages. The favorable electrical environment in which a direction finder operates on an airplane renders the use of balanced loop input unnecessary. As Coast Guard aircraft do not operate over regular air routes, which are marked with visual and radio beacons, the use of homing devices, or any arrangement which requires deviating a plane from her course, are not practicable.

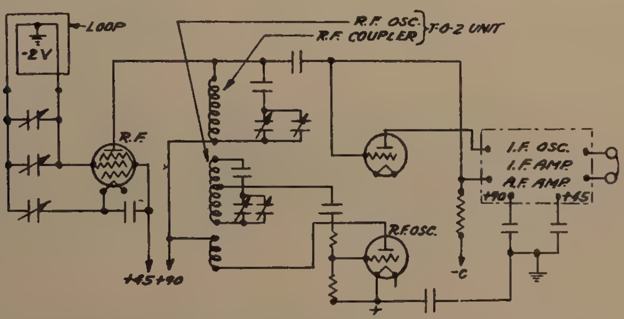


Fig. 28—Schematic circuit arrangement of Model CGR-19-B aircraft direction finder.

It is necessary for the pilot of a Coast Guard airplane to know his exact position constantly while flying on various courses along the coast and over water. A direction finder, to be of practical value under these conditions, must be located on the airplane and capable of utilizing the signals from marine radio beacons and transmitters located along the coast. By means of simple triangulation the aircraft's position can be frequently and quickly determined by means of cross bearings taken on these transmitters. To meet this requirement the rotable loop-type direction finder was selected as the most practicable arrangement. The Model CGR-19-B equipment shown in Figs. 26 and 27 has a total weight of less than 45 pounds, is readily installed in an aircraft, and requires a minimum of space. The mechanical construction throughout provides the strength necessary to insure intactness under the conditions of vibration and shock incident to operating the equipment in aircraft. The variable condensers employ steel plates in order to preserve alignment. The vacuum tubes employed are of the new low filament voltage type.

Fig. 29 shows the loop as mounted on a Navy type PN-12 seaplane. Figs. 30 and 31 show the loop as mounted on a Coast Guard amphibian plane of the "Douglas" type. The contrasting nature of the deviations

Fig. 29—U. S. Coast Guard Model CGR-19-B, radio direction finder loop mounted on U. S. Navy PN-12 seaplane.

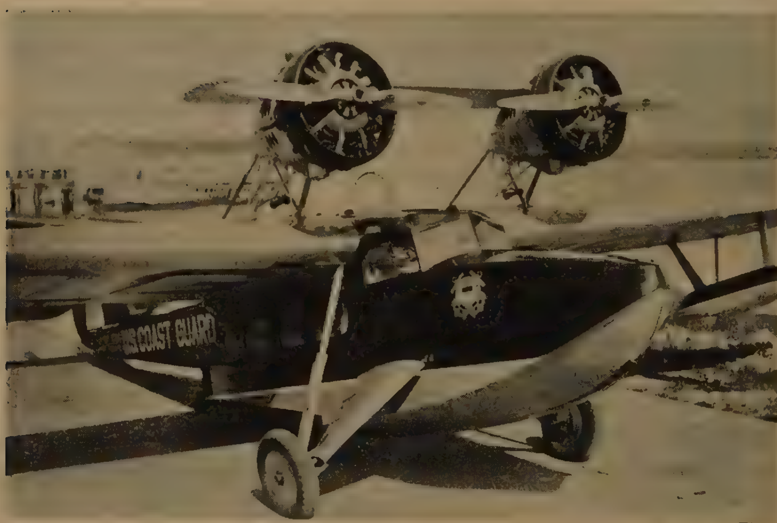


Fig. 30—Model CGR-19-B direction finder loop mounted on U. S. Coast Guard amphibian plane.

obtained are indicated by the curves in Fig. 33. The magnitude of over 20 degrees obtained on the PN-12 is due to the structural features embodied in the wings and associated equipment which provide numerous closed metallic loops arranged in various planes with regard to the ship's fore and aft axis. The overcompensated condition indicated by the curve obtained on the amphibian plane is probably the resultant of

Fig. 31—Model CGR-19-B direction finder loop mounted on U. S. Coast Guard plane.

two distinct quadrantal displacements, one resulting from the metallic hull, the other from the wing, engines, engine mounts, and landing gear. It was at first thought this reversed deviation was due to loop effect resulting from the continuity formed by the engines, engine mounts, the hull and the auxiliary wing supported by the engines. This assumption proved erroneous when the auxiliary wing was discovered to be constructed of nonmetallic materials. The deviation curves shown were obtained with the planes floating on the water and were found to hold true for bearings taken in the air up to an altitude of 1500 feet. The minima in both cases was inherently sharp and well defined throughout the entire azimuth which would be expected due to the absence of near-by vertical objects capable of acting as vertical antenna. It is believed that tests now being carried out at various altitudes up to 7000 feet will reveal interesting wave front phenomena. Bearings taken with the equipment described in this article proved consistently

accurate and compared favorably with the average results obtained with standard marine equipment. Accurate bearings on high powered coastal radio stations, such as Tuckerton, N. J., and Chatham, Mass., can be obtained up to approximately 200 miles.

The minima on all signals observed in the air was found to be exceptionally sharp and well defined. During the course of these investigations it was noted that rotating the direction finder loop in a horizontal plane while in the air produced maxima and minima which were characteristically the same as those obtained when rotating the loop in a vertical plane. It has been established by R. Keen and others that a loop located on the ground and arranged to rotate about its horizontal axis will not indicate the zenithal angle at which a signal from an aircraft is arriving. Preliminary observations made with a loop arranged to rotate in the horizontal plane in a Coast Guard aircraft indicate that an arrangement of this sort might possibly be employed for determining wave front tilt.

Fig. 32—Interior arrangement of Model CGR-19-B radio direction finder installed on U. S. Coast Guard plane.

VII. Conclusions

The following important facts have been established as a result of the conclusions drawn from an analysis of the results obtained over a

period of several years with loop-type marine radio direction finding equipment installed on vessels of the Coast Guard:

1. An average accuracy of within two degrees on radio bearings taken in the bow and stern quadrants and two to four degrees in the beam quadrants is attainable at distances up to 50 miles with signals of normal strength.

2. Reliable results can be obtained with marine equipment on low, intermediate, and high frequencies (250 to 12,000 kilocycles) with

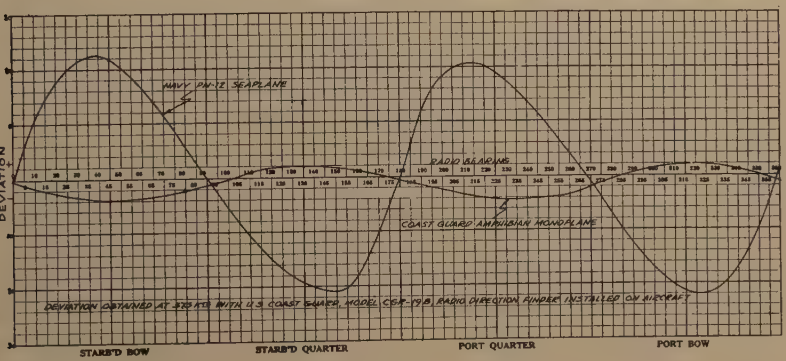


Fig. 33.

strong ground wave signals. The sky component alone, or simultaneous reception of sky and ground components, is not satisfactory for practical marine radio direction finding purposes.

3. Deviation varies with the frequency.

4. The physical dimensions of a direction finder loop have a marked effect on deviation.

5. Shielding the loop greatly reduces its susceptibility to local distortional influences.

6. A balancer arrangement of proper value will eradicate residual signal without tuning the direction finder loop or displacing the minima.

7. Quadrantal error is primarily due to the distortional influence of a vessel's or airplane's hull but is amplified by objects located thereon which, due to their electrophysical nature, can act as closed loops.

8. Obscure minima are caused by objects located near the direction finder which act as vertical antenna.

9. Many desirable advantages are gained by placing direction finders on vessels and aircraft in preference to locating this equipment on shore.

10. Excessive deviations and obscure minima, particularly in the

beam sectors, can be satisfactorily corrected by means of compensating loops.

11. Variable errors due to diurnal aberration or so-called "night effect" are frequently reported at shore radio direction finder stations while vessels in the same locality equipped with standard, Coast Guard, shielded-loop radio direction finders note absence of the effect.

12. A rotatable-doublet or "Adcock" antenna arrangement for the purpose of eliminating "night errors" is not deemed practicable for marine use, particularly at the higher frequencies, due to insufficient sensitivity. To obtain "pick-up" comparable to that realized with the usual rotatable loop, an Adcock antenna arrangement would exceed reasonable dimensions. Furthermore night effect is so infrequently experienced with a shielded and well-balanced rotatable loop that the application of any other arrangement is not deemed worthy of serious consideration.

13. The rotatable-loop direction finder is the most practicable type for aircraft which do not operate over regular airways marked with direction beacon transmitters.

14. The displacement error noted with an aircraft radio direction finder at altitudes up to 1500 feet is the same as that noted when the aircraft is resting on the surface of the water.

15. Balancing out residual signal becomes increasingly difficult and critical at the higher frequencies. The minima, however, of signals above 3000 kilocycles is so inherently sharp and well defined that balancing is not required unless the local conditions under which the direction finder is operated are exceptionally adverse.

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QUARTZ PLATE MOUNTINGS AND TEMPERATURE CONTROL FOR PIEZO OSCILLATORS*

By

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Summary—*In this paper are described a number of representative types of mountings for rectangular and circular quartz plates to be used as frequency standards. Unless the movement of the quartz plate in the holder is restricted, the frequency will change with each slight jar. A satisfactory holder for mounting a long rectangular quartz plate to vibrate at its extensional mode may be made by clamping the plate along a central line perpendicular to its length between two keys, one in the face of each electrode. The electrodes are spaced by quartz washers. A plate mounted in such a holder will be constant in frequency to 1 part in 300,000. Such a mounting has not been found satisfactory for frequencies above 100 kc as the damping caused by the pressure of the keys is too great.*

A very satisfactory holder for mounting a cylindrical quartz plate for "thickness oscillation" may be made by clamping the plate between three screws, mounted radially 120 degrees apart in a ring so that they press into a V-shaped groove cut around the cylindrical surface of the quartz plate midway between the faces. The electrodes are spaced on either side of the quartz plate by pyrex washers. Mounted in such a way, the plate has been found to be constant in frequency to 1 part in 1,000,000 in a portable frequency standard with the addition of temperature control to the oscillating circuit. Some discussion is given to the subject of temperature control of the piezo oscillator.

I. Introduction

During the past few years an extensive study of certain phases of piezo-oscillator construction and performance has been made at the National Bureau of Standards. The piezo oscillators have come from two chief sources: those sent in for calibration and those built at the Bureau to serve as frequency standards. There has been no opportunity for long-continued observations on the former class of piezo oscillators, but the latter have yielded information of considerable value based on measurements which were made daily over a period of a year. These piezo oscillators differed in the manner of temperature control, method of mounting the quartz plate, type and number of tubes used, and type of oscillating circuit. Each of these factors affects the frequency constancy of the piezo oscillators. The material presented herewith deals chiefly with the quartz-plate mounting and the temperature-control units, and is given with the hope that

* Decimal classification: R214. Original manuscript received by the Institute, September 17, 1931. Publication approved by the Director of the Bureau of Standards of the U. S. Department of Commerce. Published in *Bureau of Standards Journal of Research*, 7, 683; October, 1931.

it may be of assistance to those interested in constructing piezo oscillators of greater reliability or in judging the probable merits of a given piezo oscillator.

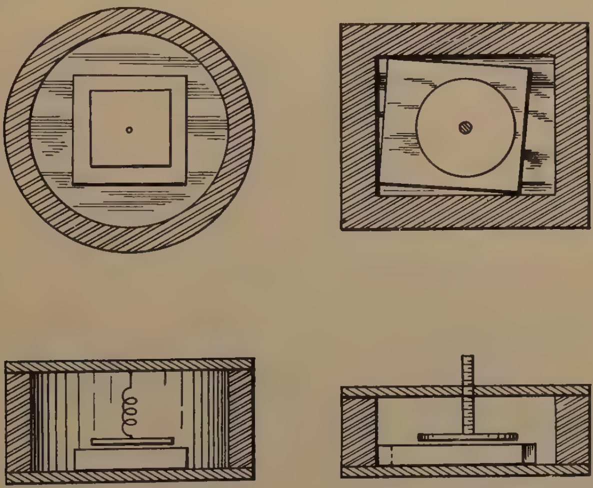
II. The Quartz Plate and its Mounting

At the present time piezo oscillators are in use which range in frequency from 25 to 5000 kc. The quartz plates commonly used in covering this range of frequencies are rectangular bars, circular disks, and thin rectangular plates. Mountings for these plates differ greatly in outward appearance but the principles involved are much the same. The mounting must be so made that: (1) the air gap will not change with aging of the component parts; (2) the air gap changes in a definite, predictable manner with variation in temperature; and (3) the unessential electrical capacity is minimized. The freedom of the quartz plate to move about between the electrodes must be restricted as much as possible without damping the desired mechanical vibration of the quartz plate. Any forces which tend to damp the vibration should be constant. The particular use for which the piezo oscillator is intended should determine the type of mounting. If the piezo oscillator is to be used as a primary standard, it will probably be kept in one position, protected from shock as far as possible, and be in continuous operation so that its frequency may be measured in terms of standard time. If the piezo oscillator is to be used as a secondary standard, it should be built so that it may be moved from place to place wherever it is needed and maintain a constant frequency over relatively short intervals of time. The frequency will change abruptly with a shift in the position of the plate within the mounting, and will drift more or less slowly with changes in damping depending on the magnitude of and rate at which the damping changes. It is very difficult, if not impossible, to maintain a constant damping factor for a quartz plate holder for an extended period of time. Therefore, it would appear that primary standards should use a mounting which possesses a minimum amount of damping, maintained, of course, as constant as possible. For secondary standards, the type of holder which allows a minimum of shifting in the position of the quartz plate is to be preferred.

The usual commercial piezo oscillator utilizes a rectangular quartz plate. The extensional mode of oscillation is customarily used for frequencies below 200 to 400 kc. For frequencies above this limit, the "thickness mode" is employed. These quartz plates are usually "Y-cut" or "30 degree-cut" quartz plates.[1]

[1] Walter G. Cady, "Piezo-electric terminology," Proc. I.R.E., **18**, 2136; December, 1930.

A simple type of holder, which is often used for rectangular quartz plates, consists essentially of two brass electrodes with a celluloid or bakelite ring to limit the shifting of the quartz plate.[2] The upper electrode may be adjustable as to air gap or may rest on the quartz plate. (See Figs. 1A and 1B). In another holder are used pins of bakelite or hard rubber mounted in the lower electrode to hold the quartz plate in place. The frequencies of piezo oscillators that incorporate mountings of this type change considerably when jarred, due to the shifting



1 a 1 b

Fig. 1—Types of simple quartz plate holder.

of the plate between the electrodes, and to variation of the damping as the plate touches the pins or a supporting ring at different points. The constancy of frequency for a quartz plate mounted in this way is ordinarily 1 part in 10,000 to 30,000.

If the two faces of the quartz plate are cleaned thoroughly, platinum electrodes may be sputtered directly on them. Connection to each electrode may be made by a piece of tin foil which is held in place by amberoid cement. The quartz plate may then be suspended by threads in a suitable position. A plate mounted in this manner will oscillate readily, but the frequency will drift continuously, usually toward a higher value. The sputtering on the surface of the plate changes the decrement and frequency of the plate. A quartz plate mounted in such

[2] A. Hund, "Uses and possibilities of piezoelectric oscillators," Proc. I.R.E., 14, August, 1926. R. C. Hitchcock, "Mounting quartz oscillator crystals," Proc. I.R.E., 15, 902; November, 1927.

a way gives a constancy of frequency of 1 part in 20,000 during a period of two or three months.

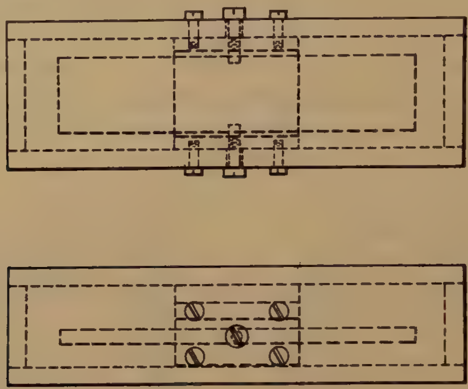


Fig. 2—Mounting for long rectangular quartz plates.

For rectangular plates which have one dimension large with respect to the other two, generally termed bars, there are other methods

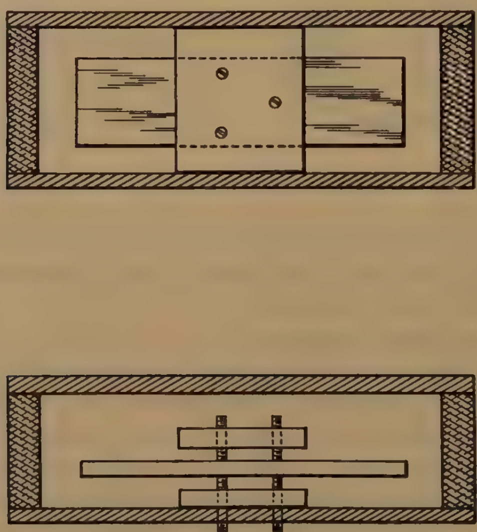


Fig. 3—Mounting for long rectangular quartz plates.

of mounting which may be used. In order to hold the quartz plate in position, a slot may be made in each edge of the quartz plate, midway between the two ends. These points are approximately at rest when the

plate is oscillating at its extensional mode. Two screws are mounted on the lower electrode in such a way that they engage in the slots, and thus hold the quartz plate in position as it rests on the lower electrode. (See Fig. 2.) The screws are made to fit the slots very nearly, and are adjusted as tightly as possible without producing binding at any normal temperature. If the electrodes cover only the central portion of the quartz plate the damping is considerably reduced. A quartz plate mounted on such a holder will be constant in frequency to 1 part in 300,000.

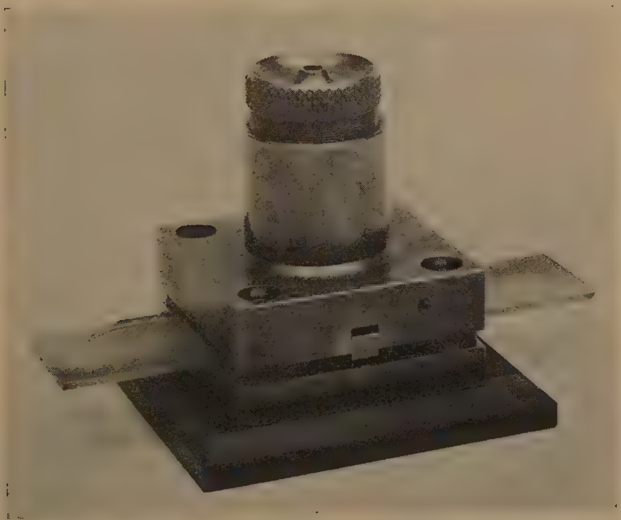


Fig. 4—Mounting for bar-shaped quartz plates.

In another piezo oscillator tested, the quartz bar is clamped between screws that project through the electrodes. The screws are located at the vertexes of an equilateral triangle; three screws in each electrode are so placed that each screw is directly opposite a screw in the other electrode. (See Fig. 3.) The position of the points of support with respect to the quartz plate must be such that a line drawn through the geometrical center of the plate, parallel to the faces and perpendicular to the length dimension, is equally distant from each screw. Any variation in temperature causes a variation in the pressure of the screws and thus changes the damping of the plate and its frequency. Even when the temperature is held constant there is often a drift in frequency. This type of mounting has proved to be unsatisfactory for a frequency standard.

Another suitable method of mounting a rectangular quartz plate

for oscillation at the extensional mode is in a holder having a narrow raised portion across the middle of the lower electrode, and a flat key directly opposite in the upper electrode, which is pressed against the quartz plate by a spring. (See Fig. 4.) The spring in the upper electrode is a stiff phosphor-bronze helix, which gives a small change in pressure with changes in temperature. Quartz spacers fix the air gap

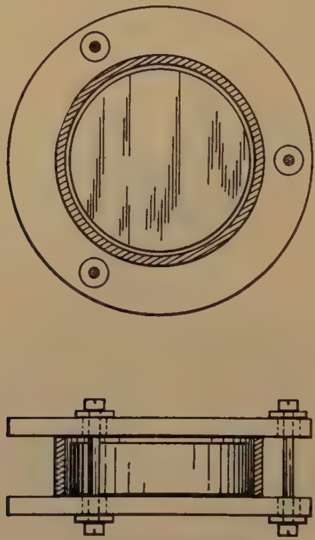


Fig. 5—Simple mounting for cylindrical quartz plates.

between the electrodes. This type of plate holder has been found very satisfactory for quartz plates whose extensional vibration frequencies are below 100 kc. Quartz plates giving higher frequencies seem to be damped too much by the width of the pressing surfaces on the electrodes and are not so satisfactory as frequency standards. A plate mounted in such a holder has been found constant in frequency to 1 part in 300,000.

A circular quartz plate is somewhat easier to mount satisfactorily than a rectangular plate. A simple type of mounting for a circular plate utilizes a piece of pyrex tubing, slightly larger in diameter than the quartz plate, to serve as the spacer between the electrodes as well as to hold the quartz plate in place.[3] (See Fig. 5.) This type of mounting is found more advantageous for a "thickness mode" than for any other mode of oscillation as there is less variation in damping for this

[3] V. E. Heaton and W. H. Brattain, "Design of a portable temperature-controlled piezo oscillator," Proc. I.R.E., **18**, 1239; July, 1930; *Bureau of Standards Jour. of Research*, **4**, 345; March, 1930.

mode. The pyrex ring must be of such a size that it will not clamp the quartz plate at any temperature at which the piezo oscillator will be operated. The greatest possible constancy of frequency of a plate in such a holder is 1 part in 200,000 for a portable standard as the quartz plate is able to shift slightly in the plate holder as well as to rotate.

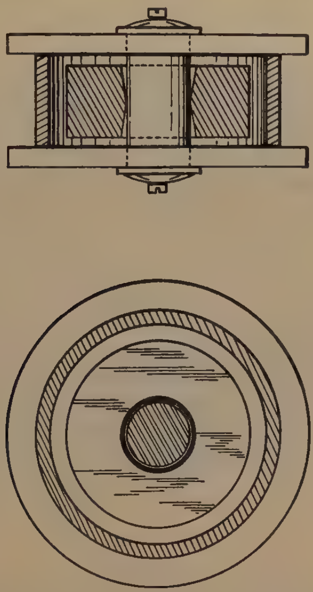


Fig. 6—A good quartz plate mounting for a stationary frequency standard.

Another dependable method of mounting a circular quartz plate involves the cutting of a tapered hole along the axis of a "30 degree" or "Y-cut" disk. The hole is cut so that the diameter decreases from both faces toward the center. The damping is small when the quartz plate is mounted on a horizontal rod of bakelite, or similar material, passed through the hole.[4] (See Fig. 6.) It has been found that the temperature coefficient of frequency may be greatly reduced for such" doughnut-shaped" plates by suitably proportioning the various dimensions. A pyrex ring serves as spacer for the electrodes. If the quartz plate moves so as to come in contact with either electrode, the decrement is increased. Therefore, such a mounting is not reliable as a portable standard, although in a fixed position the piezo oscillator is reliable to 1 part in 1,000,000 with the extension of temperature control to the oscillator circuit. If the temperature and barometric pressure of the

[4] W. A. Marrison, "A high precision standard of frequency," Proc. I.R.E., **17,** 1103; July, 1929; *Bell System Tech. Jour.*, July, 1929.

air surrounding the quartz plate and if filament and plate voltages of the oscillator tube are maintained very accurately at a constant value, the accuracy of such a piezo oscillator is 1 part in 10,000,000 according to Marrison. Experience obtained at the Bureau with similar piezo oscillators verifies this statement.

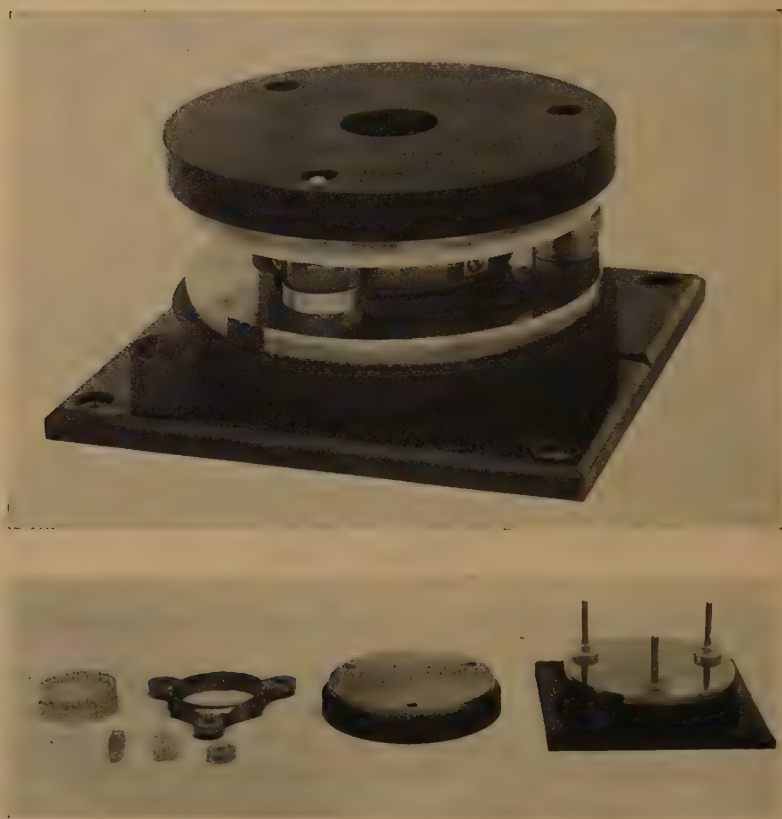


Fig. 7(a and b)—Mounting for cylindrical quartz plate (assembled and disassembled).

It is possible to mount a cylindrical quartz plate so that it will be reliable as a portable standard. About the cylindrical surface a V-shaped groove is cut halfway between the two plane faces of the quartz plate. The quartz plate is then mounted within a ring of metal by three screws, 120 degrees apart, whose tapering points fit in the groove in the quartz plate so that it is gripped only at the bottom of the groove. (See Fig. 7.) Some means must be provided for equalizing the expansion of the mounting and of the quartz plate. This may be done by

choosing a metal having approximately the same temperature coefficient as quartz or by using one metal for the ring and another for the mounting screws so that the differential expansion of the two metals will be the same as the expansion of the quartz plate. The electrodes are mounted on either side of the ring, pyrex washers serving to space them. In such a plate mounting the decrement is very small and there is no opportunity for the quartz plate to shift. A quartz plate mounted in this manner will hold its frequency constant to 1 part in 1,000,000 as a portable standard with the extension of temperature control to the oscillator circuit.

III. Temperature Control

The proper control of temperature in a piezo oscillator is just as important as proper mounting of the quartz plate. A thermostat, which operates heater units, is used to control the temperature of the medium surrounding the quartz plate mounting. This medium may be either liquid or air. A liquid bath is less satisfactory on account of the difficulty of keeping the liquid away from the quartz plate, and the lack of portability.[5] A large air bath increases the bulk of the piezo oscillator, but does not add excessive weight, and does increase the constancy of the temperature. The air bath in no way interferes with the oscillation of the quartz plate.

The simplest control consists of a small box of wood or of similar material, containing a thermostat, the quartz plate mounting and a lamp or electric heater. The constancy of temperature depends on the type of thermostat used and the relative locations of the various pieces in the temperature-controlled space. The room temperature will probably affect the temperature at which the quartz plate is maintained. If the thermostat and heaters are properly located, a good bimetal thermostat will hold the temperature constant to 1 degree C, whereas a mercury thermostat would probably increase the constancy by a factor of ten. The reasons for the difference are: first, a bimetal thermostat ordinarily has some parts outside the controlled space so that the heat may be conducted out, while the mercury thermostat is usually mounted entirely within this enclosure; and second, the variation of temperature necessary to cause the thermostat to go through one heat cycle is much less for the mercury thermostat than for the bimetal type. Over long periods of time the bimetal thermostat shows an aging which causes a change in the temperature it maintains.[6]

[5] R. C. Hitchcock, "Mounting quartz oscillator crystals," Proc. I.R.E., 15, 902; November, 1927.

[6] W. A. Marrison, "Thermostat design for frequency standards," Proc. I.R.E., 16, 976; July, 1928.

Mention has been made of the proper location of heaters, thermostat, and quartz plate mounting. If these three elements are located without any consideration or knowledge of probable results, the temperature may be found to vary several degrees within the so-called temperature-control cabinet, assuming that heat is applied to the thermostat and quartz plate by conduction through air alone. However, if the heater unit warms a large metallic body which carries the quartz plate mounting and the thermostat, the metallic body acts as a heat reservoir and distributor and effectively reduces the temperature variation required to actuate the thermostat, with a resulting reduction in temperature change of the quartz plate. Such a system, if suitably designed, using a bimetallic thermostat, may be expected to hold a given temperature much better than indicated above for such thermostats.

Another method of improving the temperature control of the quartz plate consists in placing the plate holder inside a heat-attenuating box. This box may consist of a single layer of metal, wood, or other material. It is better, however, to make the compartment of alternate layers of metal and felt or asbestos, using metal as the outside layer. The metal should be brass, copper, or aluminum so as to distribute the heat readily over the entire surface of the attenuator. The poor heat conductor between the metal walls slows down the heat transfer to the plate holder.[7] Such an attenuating cabinet of 3 or 5 layers of material may reduce the variations of temperature inside to 1 per cent of that in the temperature-controlled compartment.

Still better control may be obtained if the thermostat is placed in a well in the outer layer of metal in the attenuation box. Variations of room temperature then have a negligible effect on the temperature of the quartz plate as long as the heaters are not overtaxed.

If portability is not required, the temperature-controlled compartment may be made larger with a fan to force the air to circulate about the attenuating cabinet containing the plate holder. This arrangement is very little superior to the smaller box described above in which heat is supplied on all six sides of the attenuating cabinet.

In any temperature-control system, the thermostat operates more satisfactorily if it controls the heater current through the agency of a relay rather than directly. Either direct or alternating current may be used for the heater current. If direct current is used, a potential divider will supply the necessary current for operating the relay. If

[7] V. E. Heaton and W. H. Brattain, "Design of a portable temperature-controlled piezo oscillator," Proc. I.R.E., **18**, 1239; July, 1930; *Bureau of Standards Jour. of Research*, **4**, 345; March, 1930; J. W. Horton and W. A. Marrison, "Precision determination of frequency," Proc. I.R.E., **16**, 137; February, 1928.

alternating current is used, the relay may be operated on the output of a small rectifier.

Where great constancy of frequency is required, the oscillator circuit also must be maintained at a constant temperature. It is very desirable to have a double temperature control on the quartz plate, putting the oscillator circuit in the outer temperature-control compartment.[8] The quartz plate is mounted in an attenuating box in the outer wall of which is placed a mercury thermostat. This attenuator is in a constant-temperature compartment whose walls are of alternate layers of metal and insulating material and whose temperature is controlled by the thermostat in the wall of the quartz plate attenuator. This temperature-control unit is then placed within another compartment, whose temperature is controlled by a thermostat of either the mercury or bimetal type. This outer compartment has walls of insulating material. In the outer compartment are placed the oscillator tube, and coil. The inner compartment operates at a temperature a few degrees above that of the outer one. The other compartment contains enough heaters to maintain a constant temperature, regardless of room temperature. Such a unit will maintain a constant temperature at the quartz plate under all conditions except for small drifts in the operating temperature of the thermostat.

[8] J. K. Clapp, "Temperature control for frequency standards," Proc. I.R.E., **18,** 2003; December, 1930.

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AN IMPROVED AUDIO-FREQUENCY GENERATOR*

By

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Summary—*This paper describes in detail the construction of an audio-frequency generator, for use in making radio-frequency measurments. The variable audio-frequency output is the beat note between two sources of radio frequency; the one a piezo oscillator, and the other a variable oscillator. The output is continuously variable from 50 to 1500 cycles per second. The entire unit is assembled very compactly and the essential parts are mounted in a temperature-controlled compartment. The calibration curve is practically linear over a range of 50 cycles per second and repeated calibrations indicate that it is constant to better than 0.1 cycle per second over the entire range.*

An AUDIO-FREQUENCY oscillator is an essential part of radio-frequency measuring equipment. The last step in such a measurement is the determination of the frequency of the audio beat note between the reference standard and the unknown frequency. If the frequency of the beat note is between 0 and 200 cycles per second, an audio-frequency bridge[1] provides a very accurate means of making the measurement. However, if the frequency is greater than 200 cycles per second, the most convenient method is to match the note with a known audio frequency. A variable audio-frequency generator is ordinarily used for this purpose.

The beat-frequency oscillator[2] provides a very convenient audio-frequency source. The audio-frequency note is the beat between two radio-frequency oscillators: a piezo oscillator and a variable oscillator. The beat note is received by a detector and amplified to the desired energy level by means of an audio-frequency amplifier.

A very satisfactory audio-frequency generator of this type has been constructed for use in measuring the standard frequency transmissions from the National Bureau of Standards radio station, WWV. The complete unit can be seen in Fig. 1. It is a very compact unit, mounted in an aluminum case, $18\frac{3}{4}$ inches long, $12\frac{3}{8}$ inches high, and $12\frac{3}{4}$ inches

* Decimal classification: R355.9. Original manuscript received by the Institute, September 19, 1931. Publication approved by the Director of the Bureau of Standards of the U. S. Department of Commerce. Published in *Bureau of Standards Journal of Research*, 7, 691; October, 1931.

[1] N. P. Case, "A precise and rapid method of measuring frequencies from 5 to 200 cycles per second," *Bureau of Standards Journal of Research*, 5, August, 1930. Research Paper No. 195.

[2] A. Hund, "Piezo-electric generator for audio frequencies," *Bureau of Standards Journal of Research*, 2, February, 1929. Research Paper No. 40.

deep. This case is rigidly constructed of 3/16-inch sheet aluminum with all joints fitted neatly to provide an effective radio-frequency shield. In the upper left-hand corner of the front panel is the filament voltmeter and rheostat. At the right of the rheostat is a light which indicates the operation of the heater. The meter in the upper center indicates the current in the piezo oscillator plate circuit, and the one in the upper right, the plate voltage applied to the oscillator tubes. A

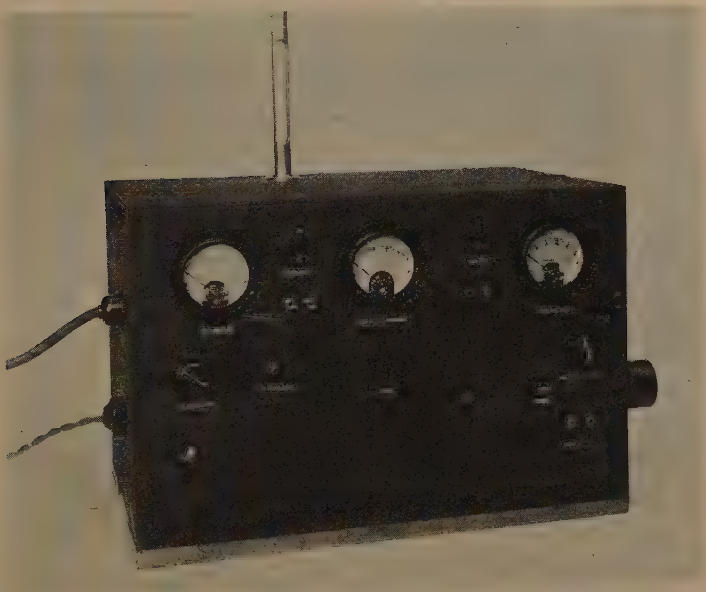


Fig. 1—Front view of audio-frequency generator.

small toggle switch is seen on either side of the plate milliammeter. The one on the left is connected in the plate circuit of the piezo oscillator, and the one on the right in the plate circuit of the variable oscillator. The outlets directly below the switches provide the corresponding radio-frequency outputs, by means of which either oscillator can be measured separately against a radio-frequency standard, or the piezo oscillator can be used as a secondary standard. A switch, directly below the plate milliammeter, disconnects the filament battery. The audio-frequency output terminals and volume control are located in the lower right-hand corner of the front panel. In the extreme left a door is provided which gives access to a compartment containing a relay, rectifier, and resistors in the heater circuits. The battery cable and the 110-volt a-c line are connected by means of suitable outlets at the left end of the box.

By unsoldering seven leads that are connected to the circuits in the temperature-controlled compartment, and removing as many screws,

Fig. 2—View of audio-frequency generator with front panel removed.

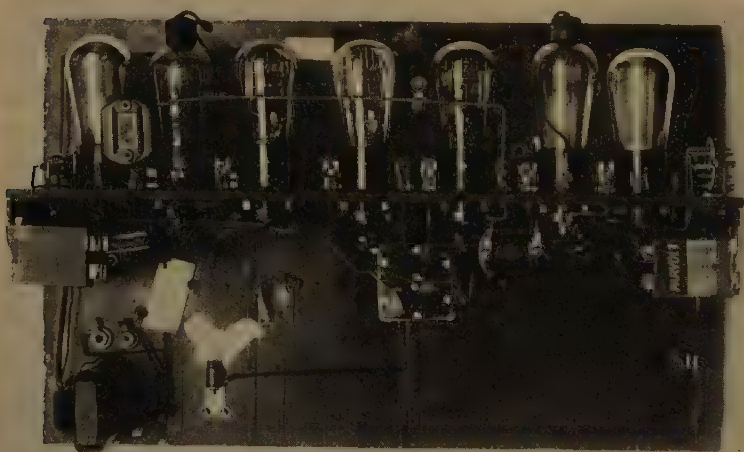


Fig. 3—Tubes and circuit arrangement as seen from rear of front panel.

the front panel may be removed. The heater control equipment can be seen in Fig. 2 at the left. On the right is a condenser dial with a

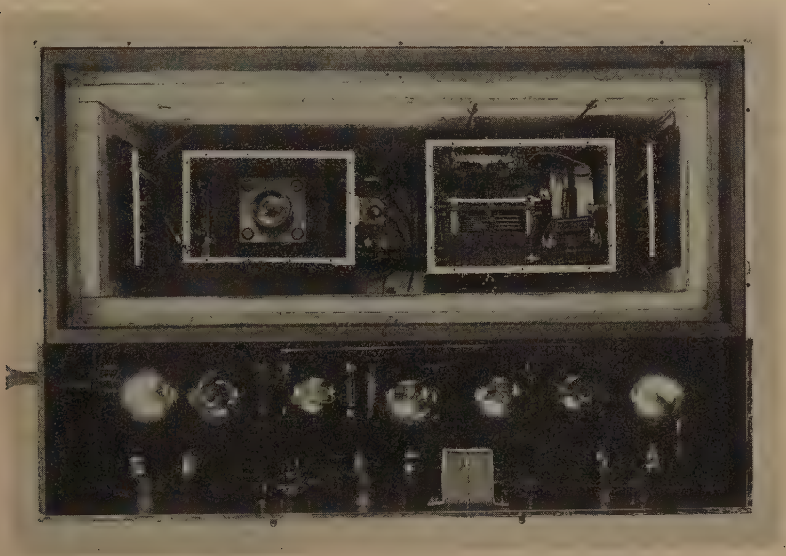


Fig. 4—Interior of temperature-controlled cabinet.

vernier, which allows a slow, smooth adjustment of the frequency of the variable oscillator.

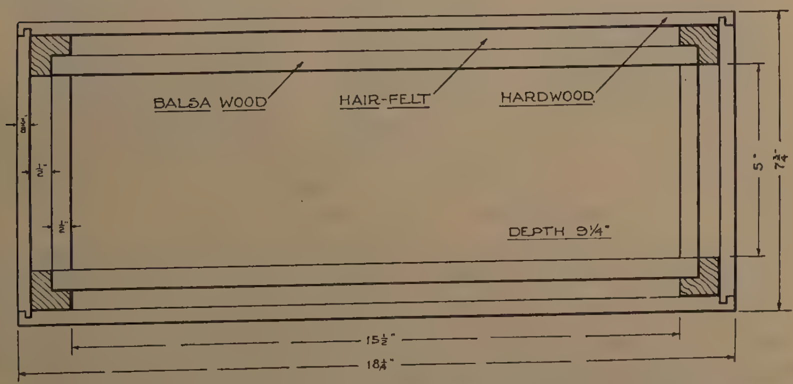


Fig. 5—Plan of temperature-controlled cabinet, showing layers of hair-felt and balsa wood.

Fig. 3 shows the arrangement of tubes and circuits as seen from the rear of the front panel. The two tubes on the right are the oscillator and coupling amplifier in the piezo-oscillator circuits. The three tubes in

the center are, from right to left, detector, first and second audio-frequency amplifiers. The two tubes on the left are the oscillator and coupling amplifier in the variable oscillator circuits.

Replacing the front panel and removing the top gives a view, as seen in Fig. 4, of the interior of the temperature-controlled cabinet. This compartment is 18¼ inches long, 12 inches high, and 7¾ inches deep, outside dimensions. The side and end walls, as shown in the drawing (Fig. 5), consist of a ⅜-inch layer of hard wood, ½-inch layer of

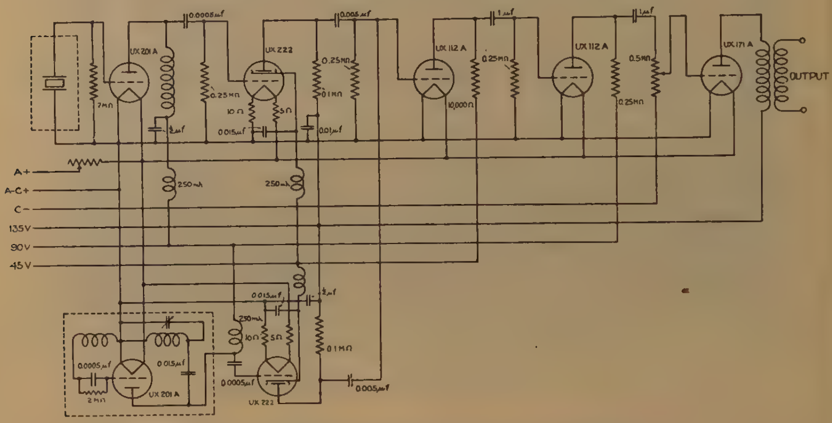


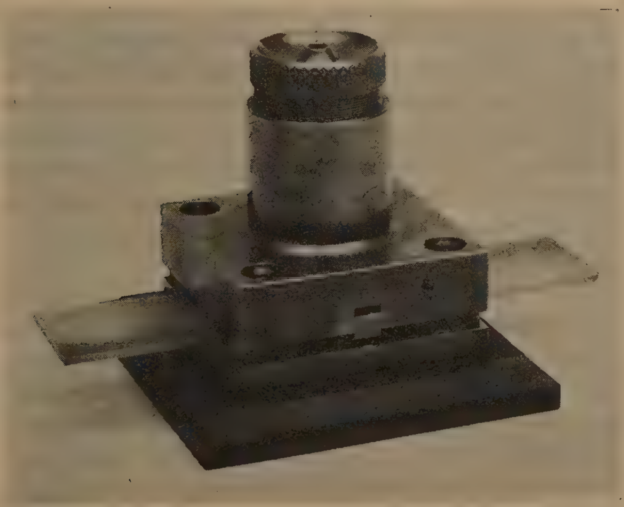
Fig. 6—Complete circuit diagram of audio generator.

hair-felt, and an inner wall of ½-inch balsa wood. The top and bottom of the cabinet consist of an outer wall of ⅜-inch hardwood and a double layer of ½-inch balsa wood.

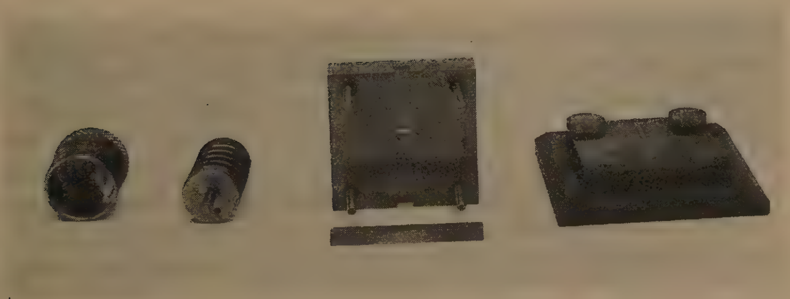
The quartz plate and mounting can be seen in the aluminum attenuating box at the left end of the temperature-controlled cabinet. Directly in front, mounted on the balsa wood, is the coil in the piezo-oscillator plate circuit. An attenuating cover is placed over this coil to reduce the temperature changes which take place as the heaters operate. The aluminum attenuating box on the right contains the condensers and coils of the variable radio-frequency oscillator. The flexible connection on the condenser shaft is in view between the front of the aluminum box and the balsa wood wall. One of the heaters, which maintain the box at a constant temperature, can be seen at each end of the cabinet. Similar heaters, extending the full length of the cabinet, are placed above and below the attenuating boxes. The thermostat is mounted between the boxes in an aluminum well directly attached to the end of the box containing the quartz plate.

The temperature-controlled cabinet is maintained at 40.5 degrees

C by means of four electric heaters. The heaters are made of No. 28 nichrome wire which is wound into a long helical spring and then



(a)



(b)

Fig. 7—Quartz plate mounting.
(a) assembled; (b) disassembled.

stretched over the forms. The wire is insulated from the metal supports by $\frac{1}{8}$-inch porcelain tubing. A sensitive polarized relay is used to operate the heaters. The relay works positively on 2 milliamperes direct current, which is obtained from the a-c line by means of a copper-oxide rectifier. This unit has been in continuous operation for four months, and in that time has not required any attention whatever.

The piezo-electric element is a 30-kc rectangular bar, 30-degree, or

Y-cut.[3] It is mounted in a special mounting, Fig. 7, in such a way that its frequency is practically free from any changes due to jarring. A spring and key in the upper electrode presses the quartz plate against a narrow, raised surface on the lower electrode. As shown in the figure, the central portion of the contacting areas are ground away so that pressure is applied only near the edges of the plate. The spring is a phosphor-bronze helix so designed that the pressure is as nearly independent of temperature as possible. In this way the quartz plate is held midway between the electrodes, and is quite free to vibrate at its extensional mode. The electrodes are spaced with doughnut-shaped (toroidal) pyrex insulators. The piezo-oscillator circuits, Fig. 6, are the same as those used in the secondary standards.[4]

The variable oscillator is a modified Hartley circuit, Fig. 6, the frequency of which can be varied from approximately 28.50 to 30.00 kc. The tuning capacity is a 250-$\mu\mu$f straight-line frequency receiving condenser connected in parallel with a 200-$\mu\mu$f mica condenser. The hard rubber insulation which was used in the variable condenser has been replaced by pyrex insulation in order that the oscillator may maintain its calibration over a longer period of time. A 15-$\mu\mu$f variable condenser is connected in parallel with the other tuning capacities in order that the audio output may be adjusted to zero cycles per second when the main tuning condenser is set on the zero position. If some other audio-frequency standard is available, the output can be adjusted to exact agreement with that point on the calibration curve instead. The adjustment for this condenser is a bakelite rod brought out through the back of the case. The range of the audio-frequency generator, which, in this case, is only from 50 to 1500 cycles per second, can be extended as desired by adjusting the capacities of the fixed and variable condensers.

The outputs of the two 30-kc oscillators are amplified separately and fed into a common detector. The resulting audio-frequency note is passed through a two-stage resistance-coupled amplifier. The resistance-capacity coupling is used in order to obtain approximately constant amplification over the entire audio-frequency range. A transformer is used in the ouput circuit to prevent any change in beat note due to a change in external connections.

The quality of the audio-frequency note which is obtained is largely dependent on the audio-frequency amplifier. If a pure sinusoidal wave form were required it is probable that some refinements in

[3] W. G. Cady, "Piezo-electric terminology," Proc. I.R.E., **18**, 2136; December, 1930.

[4] V. E. Heaton and W. H. Brattain, "Design of a portable temperature-controlled piezo oscillator," Proc. I.R.E., **18**, 345; July, 1930; *Bureau of Standards Journal of Research*, **4**, March, 1930. RP 153.

the amplifier would have to be made. However, no attempt has been made to determine the exact wave form under the present conditions. For the purpose for which the audio-frequency generator was designed, the quality of the note is entirely satisfactory.

The calibration of the audio-frequency generator has been checked repeatedly for some time. The original calibration and all recalibrations are made against harmonics of the 100-cycle-per-second output of the primary frequency standard. These measurements are made in the following manner. The condenser is set on the setting for 100 cycles per second, which is obtained from the original calibration curve. The audio-frequency note is then adjusted to exact agreement with the standard by means of the small auxiliary condenser previously described. The settings for harmonics of the 100 cycles are then determined. For any setting of the condenser the frequency has been found to vary less than 0.1 cycle per second. Furthermore, if the plate voltage is maintained at a constant value, the auxiliary condenser will not have to be readjusted for several hours at a time, which permits one to make measurements over an extended period without interruption.

Proceedings of the Institute of Radio Engineers
Volume 20, Number 2 *February, 1932*

SOLAR ACTIVITY AND RADIOTELEGRAPHY*

By L. W. Austin

(Bureau of Standards, Washington, D.C.)

Summary—*This report to the International Research Council on solar and radio relationships shows that the relationships are closer at short wavelengths than at long, that the effect of magnetic storms, which are assumed to be due to solar action, is, in general, to weaken night signals at all wavelengths and in the medium and long-wave ranges to strengthen day signals. Curves are given which show that there is a direct correlation between the yearly averages of long waves, daylight transatlantic signals, sun spot numbers, and magnetic activity (1915–1930), a direct correlation between signals and magnetic activity averages by months (1924–1930), and an inverse correlation between sun spots and atmospheric disturbances averages by years (1918–1930).*

Since the last report[16] of the Committee on Solar and Terrestrial Relationships of the International Research Council, the evidence for the reality of a connection between solar and terrestrial magnetic activity on the one hand, and radio wave propagation and atmospheric disturbances on the other, has been considerably strengthened.

Some uncertainty has been caused by the differences in the behavior of radio waves of different wavelengths under the influence of magnetic storms, but the main facts thus far ascertained are as follows:

The sensitiveness of radio transmission to magnetic storms decreases with the wavelength.[2,9,12,13] Below sixty meters, even slight storms are generally accompanied by radio disturbances. In this range of wavelengths the effect appears always to be a depression of signal strength both by day and night. The signal depression usually reaches its lowest point on the day of the greatest magnetic disturbance, with a gradual rise to normal values continuing for several days.[12] According to Maris and Hulburt,[16] when a magnetic storm starts, only those short-wave radio paths are affected which lie on the daylight side of the earth; those on the dark side remaining quiescent until dawn. This is in agreement with their theory which ascribes magnetic storms to a flash of ultra-violet light from the sun. The influence of magnetic storms on short-wave radio transmission is generally (but not always) more marked in the case of signal paths at right angles to the magnetic meridians than on those parallel to them; the effects being greater, the

* Decimal Classification: R113.5. Original manuscript received by the Institute, September, 18, 1931. Publication approved by the Director of the Bureau of Standards of the U.S. Department of Commerce.

[16] Members refer to bibliography.

nearer the paths lie to the magnetic poles. According to T. L. Eckersley,[15] the number of periods in the year from October, 1927, to October, 1928, in which magnetic storms rendered short-wave communication impossible between England and Montreal, were 49; between England and New York 32; while between England and points to the southward, in Australia, South Africa, India, and South America, these periods varied from four to seven.

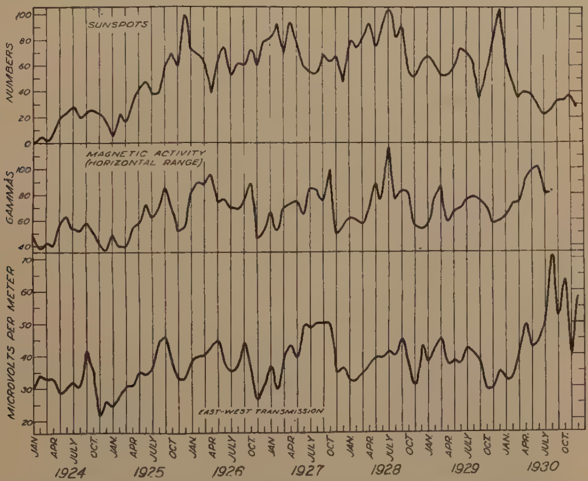


Fig. 1—Monthly averages of daylight signals from several European stations, 10,000 to 20,000 meters wavelength, measured at the Bureau of Standards from 1924 to 1930, with the monthly averages of sun spot numbers and the horizontal magnetic range.

In the range of wavelengths from 200 to 500 meters, the most noticeable effect of magnetic storms is a depression of night signals.[3,4,5] There seems to be no certain information concerning daylight effects.

At a wavelength of 5000 meters, the magnetic disturbances sometimes are accompanied, but more often are followed after a delay of one to three days, by a daylight strengthening of signals amounting to from 30 to 75 per cent, and a marked night depression with a gradual return to normal conditions.[2,9,12] At very great wavelengths, above 10,000 meters, the effects of magnetic storms on transmission become much less marked. Too few observations on European stations have been made at the Bureau of Standards during the hours of all darkness signal path to give any information regarding such relationships. The partial darkness path signals received about 3 P.M. appear to vary in general with the all-daylight signals of the morning reception. Some observations of the Bell Telephone Company,[2,9] however, have in-

dicated that the night depression effect exists at 17,000 meters. At these wavelengths the connection between the storms and the daylight signals is so slight that, with the exception of occasional instances,[11,17] it is made evident only by averaging the effects of a large number of storms. Miss Wymore[11] has shown by statistical methods that the maximum rise in Washington of very long-wave European daylight signals comes as a rule from one to three days after the height of the magnetic storm.

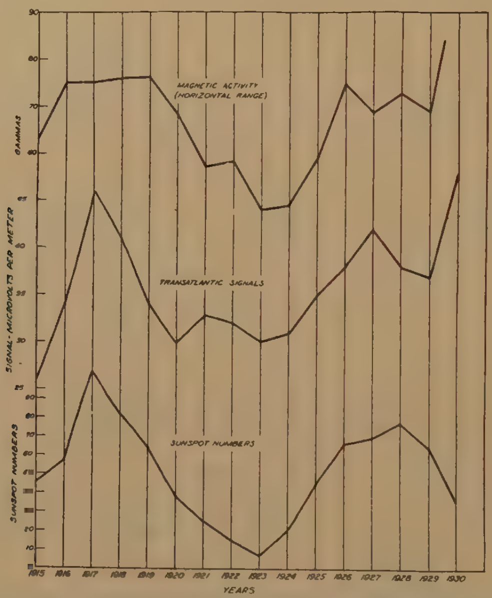


Fig. 2—Annual daylight signal averages from several European stations, 10,000 to 20,000 meters wavelength, measured at the Bureau of Standards from 1915 to 1930, with the annual averages of sun spot numbers and the horizontal magnetic range.

According to Appleton,[14] the increase in signal strength of long-wave stations at the time of a magnetic storm is due to the increase of ionization of the Kennelly-Heaviside layer at that time, the long waves being turned back to earth at its lower surface by a form of reflection which is increased as the ionization becomes greater. The short waves, on the contrary, are weakened by the increased ionization since they penetrate the layer more deeply and are subject to increased absorption.

Fig. 1 shows the monthly averages of the daylight signals from several 10,000- to 20,000-meter European stations as measured at the Bureau of Standards[18] from 1924 to 1930. The figure also shows the

monthly averages of sun spots and of the horizontal magnetic range* as measured at Cheltenham, Maryland, for the same period. It is seen that the sun spot curve does not show much resemblance to the curve of signals nor of horizontal magnetic range but that the resemblance between the two latter curves is very striking.

Fig. 2 shows the yearly daylight signal, sun spot, and horizontal

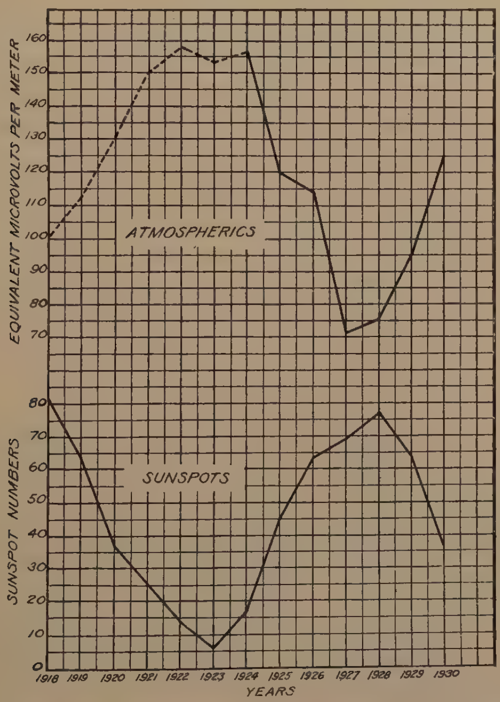


Fig. 3—Annual averages of atmospheric disturbances, 12,500 meters wavelength, at Washington, 3 P.M., Eastern Standard Time, and sun spot numbers, from 1918 to 1930.

range averages from 1915 to 1930,† inclusive. The signal measurements were taken at the Bureau of Standards [6,18] from 1915 to 1923 on the 13,000-meter station at Nauen (Berlin). During this time the shunted telephone method was used in the measurements and the accuracy is much less than in the later years. From 1924 the curve is made up from observations on a number of European stations having wavelengths between 10,000 and 20,000 meters. The great increase in strength of signal in 1930 and the increase in magnetic activity in spite of the fall

* The horizontal ranges at Cheltenham have not been calculated since August, 1930.

† The horizontal range average for 1930 is for the first half of the year.

in the number of sun spots is noteworthy. The long-wave daylight signal intensities began to rise sharply in March, 1930, reached a maximum in August and even at the present writing (February, 1931) continue to be above normal. During the summer of 1930 the short-wave communication was reported to be the weakest thus far experienced, illustrating apparently the opposite effect of magnetic activity on long daylight waves and on short waves.

Fig. 3 represents the annual averages of the 3-P.M. atmospheric disturbances measured in Washington at a wavelength of 12,500 meters, as compared with the sun spot numbers from 1918 to 1930.‡ The dotted part of the disturbance curve is less accurate than the part represented by the solid line. It appears from this that there is very probably an inverse relationship between the annual averages of long-wave daylight atmospheric disturbances and sun spot numbers when averaged in periods of a year or more. Very little success has attended the effort to show a connection between the disturbances and solar and magnetic activities when averaged in periods of less than a year, probably on account of the disturbing influences of weather variations.

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13. K. Sreenivasan, "On the relation between long wave reception and certain terrestrial and solar phenomena," Proc. I.R.E., **17**, 1793, 1929.
14. E. V. Appleton, Solar Activity and Wireless Transmission. Second Report of Committee on Solar and Terrestrial Relations of the International Research Council, p. 16, 1929.

‡ In the last report of this Committee, page 18, similar curves were shown up to 1926.

15. T. L. Eckersley, "An investigation of short waves," *Jour. I.E.E.* (London), **67**, 992, 1929.
16. L. W. Austin, Solar Activity and Radio Transmission. Second Report of Committee on Solar and Terrestrial Relations of the International Research Council, p. 18, 1929.
17. P. A. de Mars, G. W. Kenrick, and G. W. Pickard, "Low-frequency radio transmission, Proc. I.R.E., **18**, 1488, 1930.
18. L. W. Austin, E. B. Judson, and I. J. Wymore-Shiel, "Solar and magnetic activity and radio transmission," Proc. I.R.E., **18**, 1997, 1930.

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INVESTIGATIONS OF KENNELLY-HEAVISIDE LAYER HEIGHTS FOR FREQUENCIES BETWEEN 1600 AND 8650 KILOCYCLES PER SECOND*

By

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Summary—*The results of observations of the height of the Kennelly-Heaviside layer carried out near Washington, D. C., during 1930 are presented. Evidence for the existence of two layers (corresponding closely in virtual height to the E and F regions discussed by Appleton) is found during daylight on frequencies between three and five megacycles. The modification in the virtual height of the higher F layer produced by the existence of a low E layer is investigated theoretically, and the possibility of large changes in virtual height near the highest frequency returned by the E layer is pointed out. A number of oscillograms showing the characteristic types of records observed during the tests are presented together with a graph of average heights from January to October, 1930.*

I. Introduction

IT IS the purpose of this paper to describe tests conducted by the National Bureau of Standards for the study of the Kennelly-Heaviside layer subsequent to those recently reported.[1] The method employed in these tests is the group retardation or pulse method

Fig. 1a—Type of pulse received in the absence of sky waves. (Interval between adjacent timing marks in all oscillograms shown is 1/120 second.)

described by Breit and Tuve.[2] In this method pulses of about 5×10^{-4} second duration are transmitted with a group frequency of about 30

* Decimal classification: R113.61. Original manuscript received by the Institute, October 24, 1931. Publication approved by the Director of the Bureau of Standards of the U. S. Department of Commerce. Presented at annual meeting of American Section, International Scientific Radio Union, May 1, 1931. Published in *Bureau of Standards Journal of Research*, 7, 1083; December, 1931.

[1] T. R. Gilliland, "Kennelly-Heaviside layer height observations for 4045 kc and 8650 kc, *Bureau of Standards Jour. Research*, 5, (RP246), 1057, 1930; and Proc. I.R.E., 19, 114; 1931; P. A. de Mars, T. R. Gilliland, and G. W. Kenrick "Kennelly-Heaviside layer studies," Proc. I.R.E., 19, 106–113; January, 1931.

[2] G. Breit and M. A. Tuve, "A test of the existence of a conducting layer," *Phys. Rev.*, 21, 554, 1926, and Proc. I.R.E., 17, 1513, 1929.

pulses per second. The signals received at a distant point are recorded by means of an oscillograph.

Fig. 1a shows the rectified form of the pulse as received in the absence of sky waves. Received patterns showing more than one peak for a single transmitted pulse are attributed to waves arriving over

Fig. 1b—Lower trace shows pulse pattern received at 11:20 A.M., E.S.T., on August 16, 1930 (frequency 4045 kc). Note strong peaks corresponding to 100 km and 330 km virtual heights in addition to ground wave peak. Upper trace shows more complicated pattern at 11:24 A.M. Here we find strong peaks corresponding to virtual heights of 100, 290, and 330 km. Both traces read from left to right.

more than one path, that is, a ground wave and single or multiple reflections from one or more ionized strata or layers. Patterns indicating reflections are shown in Fig. 1b.

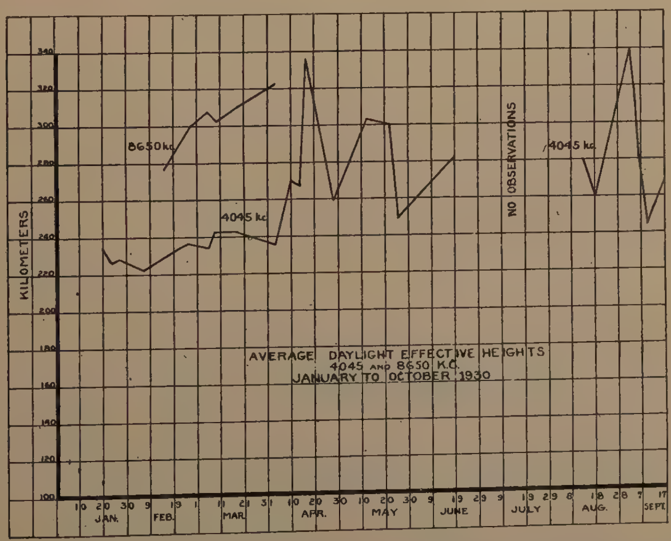

Fig. 2—Average daylight (virtual) heights of the Kennelly-Heaviside layer from January to October, 1930. (4045 and 8650 kc). Note sky waves returned on 8650 kc only between February and April of the period of observations.

The pulse method has been extensively employed in America in Kennelly-Heaviside layer studies. In England, phase interference

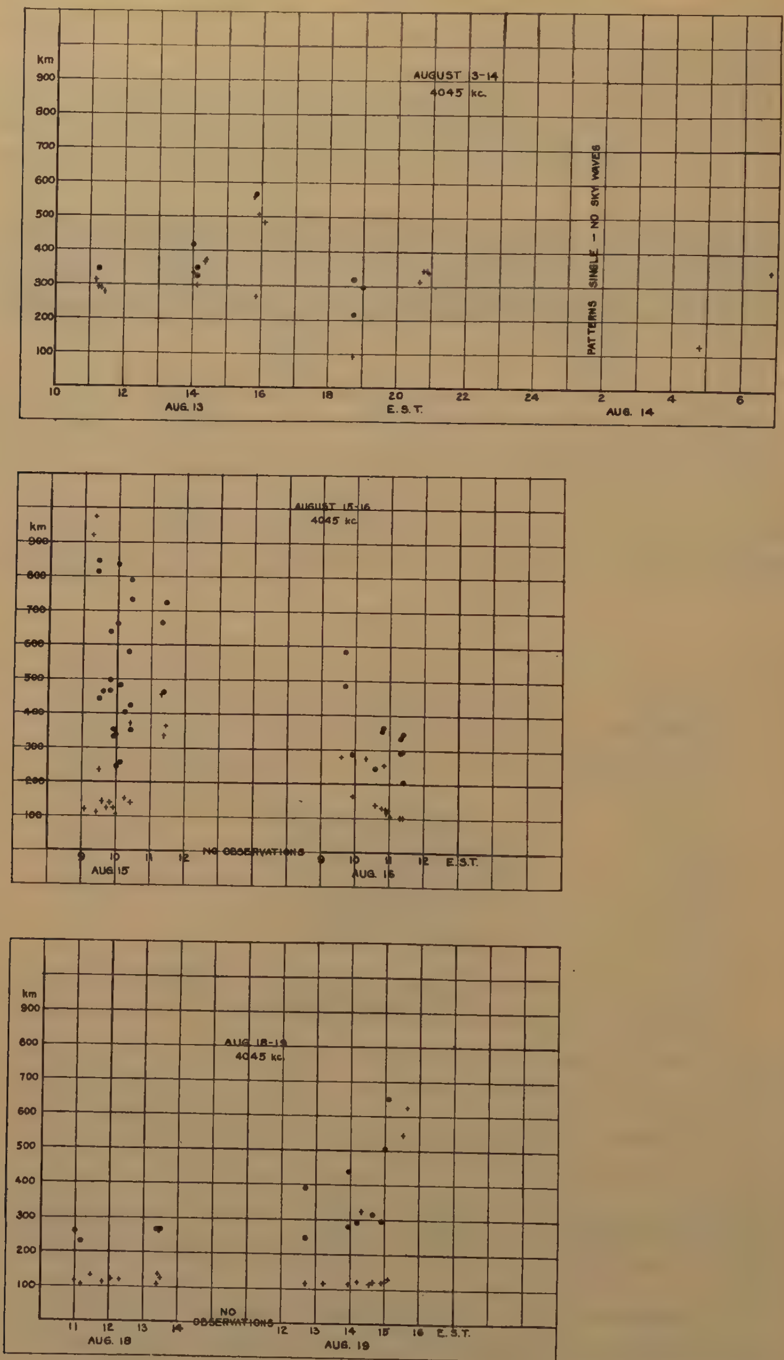

Fig. 3—Scatter diagram of virtual heights observed during period extending from August 13 to August 19, 1930. Note frequent occurrence of 100-km heights.

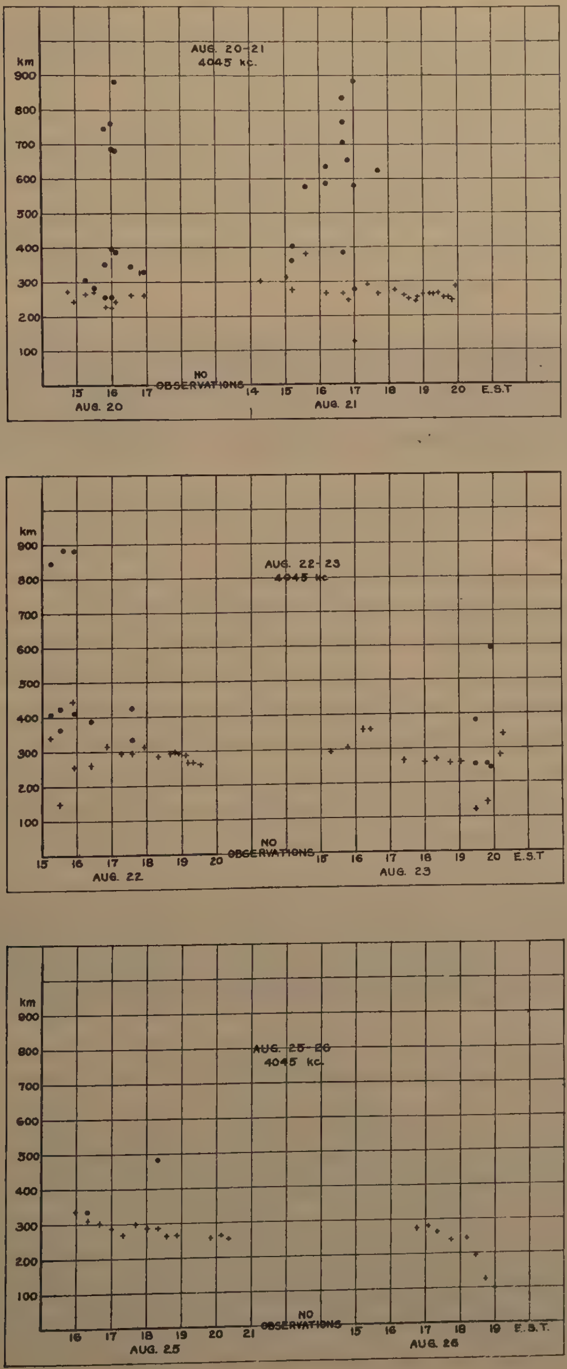

Fig. 4—Scatter diagram of virtual heights observed during the period extending from August 20 to August 26, 1930. Note rare occurrence of 100-km heights.

methods have been used by E. V. Appleton and coworkers.[3] It is of interest to note the good accord of the results obtained by the methods outlined here with those of Appleton.

Both methods have peculiar advantages, and, as has been emphasized, depend upon somewhat different analytical relations for their interpretation. It has been shown, however, that the results obtained by both methods should be in accord.[4] A brief review and comparison of the methods follows; the reader is referred to the references noted for a more extended discussion.

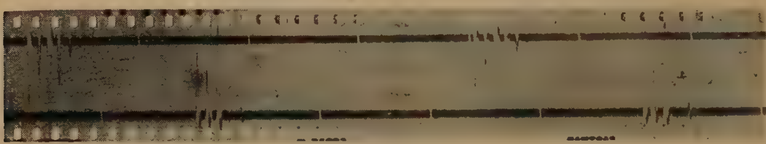

Fig. 5—Upper and lower traces both show example of multiple reflections. (4054 kc, 11:30 A.M., E.S.T., August 18, 1930).

Briefly, Appleton's method utilizes the phase reënforcements and cancellations produced by slight changes in emitted frequency due to the presence of multiple transmission paths having different phase retardations (that is, due to sky waves). The group retardation observations, on the other hand, measure the time retardation between the pulses arriving over different paths. The Appleton method employs balanced open antenna and loop antenna systems which permit measurements of polarizations and other phenomena not directly disclosed by group retardation studies when a nondirectional antenna is

Fig. 6—Note predominant 565-km virtual height. Note, however, faint evidence lower layer in upper trace. (4045 kc, 3:50 P.M., E.S.T., August 13, 1930).

employed. However, directional antennas may be used to advantage to obtain additional information when pulse methods are employed. An interesting preliminary application of pulse methods to a study of

[3] *See*, for instance, E. V. Appleton, *Proc. Royal Soc.*, **109**, 621, 1925; **113**, 450, 1926; **115**, 305, 1927; **117**, 576, 1929; **126**, 542, 1930; **128**, 134, 1930; **128**, 158, 1930. An application of a method essentially equivalent to the pulse method of Breit and Tuve has been described recently by T. L. Eckersley, "Multiple signals in short-wave transmission," Proc. I.R.E., **18**, 106; January, 1930.

[4] E. V. Appleton, *Proc. Phys. Soc.* (London), **41**, 43, 1928; also **42**, 321 June, 1930; Kenrick and Jen, Proc. I.R.E., **17**, 711–733; April, 1929.

phase relations has been described.[5] During the observations described here it was found convenient to combine the input from a horizontal and a vertical antenna in suitable proportions, in order to render the amplitudes of the ground and sky waves comparable.

The group retardation method has the advantage that sky waves arriving from several paths of widely different retardations (or from several distinct layers) are readily and immediately resolved from the oscillograms, without the use of the more elaborate harmonic analysis required in the interpretation of results obtained by phase interference methods. (See Fig. 1b.) This is of marked importance when multiple reflections and a large number of paths contribute to the results. (See Fig. 5.) Investigations of this type have emphasized that, particularly during night conditions, the number of paths may be very great indeed and far from sharply defined.[6] Pulse methods may be used to investigate transmission on relatively low frequencies, down to perhaps 200 kc per second. On lower frequencies, however, the length of the desired pulse begins to approach the period of the transmitted frequency and the time constant of the antenna. After this, difficulties introduced by transients and other obvious modulation limitations govern. The phase interference methods are, however, not free from difficulty at these lower frequencies, due to the large change of frequency required to produce the interference patterns. Under such conditions, a modification of the phase interference method, in which the path is varied instead of the frequency, is of particular interest.[8] The good accord obtained by the various methods despite the difficulties peculiar to each, is encouraging. Some comparisons of this sort are included here.

In the work described here oscillographic records were made on frequencies ranging from 1600 kc per second to 8650 kc per second (including a number not previously observed).

The observations include the summer of 1930, which was one of marked magnetic disturbance. Observations during this period are hence of particular interest, although they can hardly be considered as typical summer conditions. Of particular interest in this connection is the notable absence of multiple peaks on the higher frequencies

[5] L. R. Hafstad and M. A. Tuve, "An echo interference method for the study of radio wave paths," Proc. I.R.E., 17, 1786–1792; October, 1929.

[6] de Mars, Gilliland, and Kenrick, Proc. I.R.E., 19, 106–113, January, 1931.

[8] J. Hollingworth, "The propagation of radio waves, *Jour. I.E.E.*, (London), 579–595, May, 1926. A recent interesting application of this method has also been reported on much higher frequencies. *See* C. B. Mirick and E. R. Hentschel, Proc. I.R.E., 17, 1034–1041; June 1929. At the higher frequencies a rapidly moving recording system is essential to this method due to frequent inadvertent path changes during the observations from other causes such as layer movements, etc.

during the summer months, although they were frequently observed on 8650 kc during the winter and early spring. (See for instance the curves of Fig. 2 where the average daytime virtual heights on 4045 kc and 8650 kc are indicated.) It should be pointed out that although the

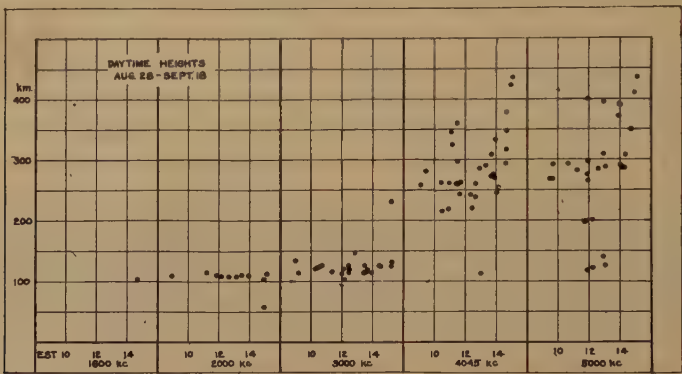

Fig. 7—Daytime heights observed for 1600, 2000, 3000, 4045, and 5000 kc between August 28 and September 18, 1930.

same transmitter was used on all of the frequencies during August and September, the radiated power was not measured and may have been enough less on the highest frequencies to indicate a cut-off frequency somewhat lower than that actually existing.

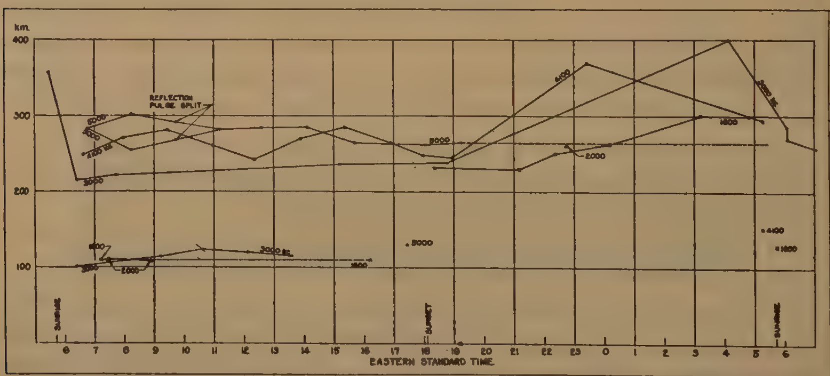

Fig. 8—Diurnal changes on 1600, 2000, 3000, 4100, and 5000 kc September 17–18, 1930. The lowest distinct height from each oscillograph record is indicated by a cross, and distinct greater heights (not in simple multiple relation) are denoted by dots.

It will be noted that these curves show only predominant virtual heights, Diagrams are, however, also included, showing the observed virtual heights throughout the 24 hours on a number of frequencies. (See Figs. 3, 4, 7 and 8.) Distinct evidence of the existence of two

sharply defined regions corresponding to the *E* and *F* regions of Appleton, is also to be noted in a number of cases. An approximate analytical investigation of the effect of a lower layer on observed virtual heights from higher layers is also indicated.

II. 4045-kc Observations

Fig. 2 presents graphs giving average daylight virtual heights on 4045 and 8650 kc, from January, 1930, to October, 1930. The observations from January to June were on transmission from the Naval Research Laboratory at Bellevue, Anacostia, D.C. These observations were interrupted when no transmitting facilities were available at the Naval Research Laboratory for this purpose during the summer of 1930. However, during the early summer a transmitter was placed in operation at the National Bureau of Standards experimental station near Alexandria, Va., which permitted the observations described to be made during August and September. All oscillograms of the received signals were recorded at the National Bureau of Standards experimental station at Kensington, Md. The distance from the transmitter to receiver was, in each case, about 20 km.

It will be noted by reference to Fig. 2 that on 8650 kc only single pulses were received, except for the period from February 6 to April 7 when multiple patterns were obtained. A rise in virtual height on 4045 kc followed the disappearance of the multiple patterns on 8650 kc. It will be noted that this high virtual height persisted during the summer with temporary fluctuations to lower values, but in all cases the pattern on 8650 kc remained single during the period of observations reported.

In addition to the very high values of the virtual height of the *F* layer, distinct evidence of the existence of the *E* layer on 4045 kc during the daytime was to be found on numerous occasions during the middle of August (Fig. 3). Fig. 1b gives an oscillograph trace showing this phenomenon, and scatter diagrams contrasting this effect with conditions a few days later are shown in Fig. 4. The actual values of virtual heights as read from the records are plotted as points, the lowest distinct height from each record being indicated by a cross, and distinct greater heights (not in simple multiple relation) are denoted by dots. The marked frequency with which the 100-km layer was observed during the daytime on 4045 kc on August 15 and 16 (Fig. 3) is, it will be noted, in marked contrast to the phenomena observable a few days later (Aug. 20 and 21) (Fig. 4). An interesting phenomenon in the form of a progressive downward movement of virtual height on 4045 kc during the afternoon observed on August 25 and also on August 26, is also shown in Fig. 4. This is one of the few cases where heights between 150

and 200 km were observable; (that is, heights intermediate between the frequently occurring 100 to 130 km and 250 to 300 km virtual heights referred to as the E and F regions or layers.[3] The pronounced low layer is clearly shown in the oscillogram of Fig. 5 and the presence of a strong peak giving a high virtual height is shown in Fig. 6.

The persistent appearance of large virtual heights not readily explicable on the basis of simple multiple reflections from the E or F layers, is emphasized in many of the oscillograms (see, for instance, Figs. 1b and 6). Phenomena of this nature must doubtless be taken into account as the theory develops, for these phenomena are apparently quite typical. A possible explanation is suggested by the theory presented in this paper.

III. Virtual Layer Heights as a Function of Frequency During Daytime

During August and September frequent observations were made on 1600, 2000, 3000, 4045, 5000, 6425, and 8650 kc, approximately the same number of observations being taken on each frequency. The results obtained during daytime (9 to 15 E.S.T.) are shown in Fig. 7.

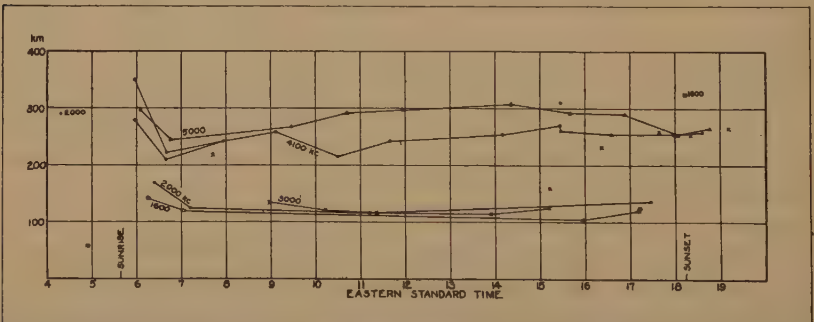

Fig. 9—Diurnal changes on 1600, 2000, 3000, 4100, and 5000 kc September 12, 1930. The lowest distinct height from each oscillograph record is indicated by a cross, and distinct greater heights (not in simple multiple relation) are denoted by dots.

No reflections were observed on 6425 and 8650 kc during these hours. This is in marked contrast to the conditions between February to April, when multiple reflections on 8650 kc were frequently observed.

The patterns obtained for many of the observations, even on the lower frequencies, showed only the ground wave. Fig. 7 indicates the number of times that reflections occurred on the various frequencies. It will be noted that virtual heights between 100 and 150 km, corresponding to the E layer, appear for 1600, 2000, and 3000 kc. Occasionally this layer was observed on 4045 and 5000 kc but virtual

[3] *Loc. cit.*

heights of over 200 km occurred most frequently on these higher frequencies. Once on 3000 kc and twice on 5000 kc, coexisting E and F layers were observed. Only the first, or in case of coexisting E and F layers, the first two reflections, were plotted in Fig. 7.

IV. Diurnal Variation

Fig. 8 shows the results obtained for frequencies of 1600, 2000, 3000, 4100, and 5000 kc for a 24-hour period (September 17 to 18).

The E layer was observed on 1600 kc shortly after sunrise and just before sunset. Otherwise no reflections were observed on this frequency during daylight. However, just after sunset the virtual height was observed to be 232 km and during the night the height increased to 300 km. At sunrise it again dropped to the E-layer region.

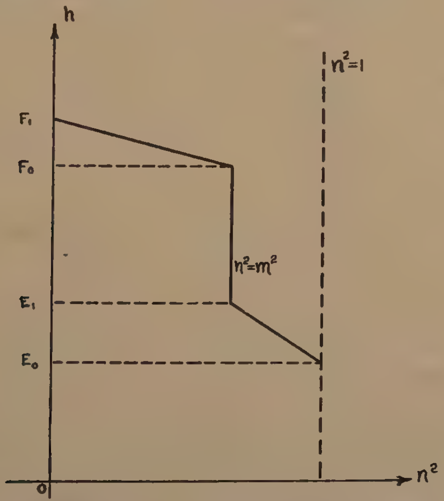

Fig. 10—Assumed change of index of refraction squared as a function of height.

On 2000 kc reflections were observable only three times during the 24 hours and the heights corresponded closely to those for 1600 kc.

Just before sunrise the height for 3000 kc was observed to be 357 km and shortly after sunrise coexisting reflections from the heights of 100 and 215 km were recorded. During the middle of the day only the E layer was observed on this frequency. After 15 E.S.T. the E layer was no longer observable but reflections from the F region were again noted.

The behavior on 4100 and 5000 kc is clearly shown on the diagram. It will be noted that on 4100 kc coexisting reflections from heights of 153 and 295 km were observed shortly before sunrise on the 18th.

Fig. 9 shows diurnal variations for the same frequencies for a slightly shorter period on September 12.

V. Analytical Discussion of Results

1. *Refraction*

The phenomena observed in these tests offer evidence in favor of the existence of at least two more or less sharply defined strata in the upper atmosphere at which refraction takes place (as already suggested by other investigations conducted previously.[3,6] The diffraction theory for one layer has already been discussed,[1,4,9] but it is of interest to consider the relations when two well-defined ionized strata exist. As Appleton has suggested, rays which are finally returned from the upper stratum nevertheless give virtual heights which may be considerably altered due to the transit through the lower layer. It is of interest to consider this possibility analytically to determine if such a view is in accord with observed phenomena. As in the case of a single layer, the theory requires an assumption as to the electron distribution in the layer, and the case of a single layer has been worked out for several such assumptions. At present it is, of course, impossible to represent the

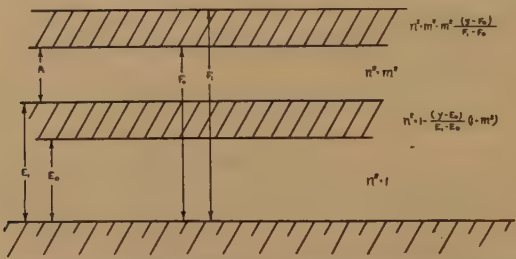

Fig. 11—Assumed layer distribution showing equations for variation of n.

exact conditions which obtain in the upper atmosphere analytically, for these conditions are very imperfectly known and involve a turbulent, rapidly varying distribution of ionization. It may be possible, however, to disclose some of the salient phenomena on the basis of simplified and necessarily idealized assumptions. In the following discussion we shall investigate such an idealized treatment to determine if it can be used to fit the facts. We shall also investigate certain other possible explanations of some of these phenomena suggested as a result of those investigations. The simplest assumption, from the standpoint of analytical development, is a distribution of electron density, such as shown in Fig. 10; that is, one which produces a linear change of the square of the index of refraction with height at the frequency under consideration. Under such conditions the path is (in the case of a single layer) parabolic. Several other assumed variations of electron

[9] P. O. Pedersen, "The Propagation of Radio Waves, published by Denmarks Naturvioenkabeliage Samfund, Copenhagen, 1928.
[1] *Loc. cit.*
[3] *Loc. cit.*
[4] *Loc. cit.*
[6] *Loc. cit.*

density giving other types of trajectories have been investigated[4] and shown to yield rather similar results in the single layer case, provided the curves do not have a minimum index of refraction near zero. Under such conditions long retardations may be encountered.[10] These distributions, however, represent critical border line conditions and will not be considered here. In view of the above considerations, and with due regard to the simplicity attainable, the linear variation of squared index of refraction of the form shown in Fig. 10 has been selected.

The postulated distribution considers a linear decrease in the square of the index of refraction starting at a height E_0 and decreasing linearly to a value m^2 at a height E_1.

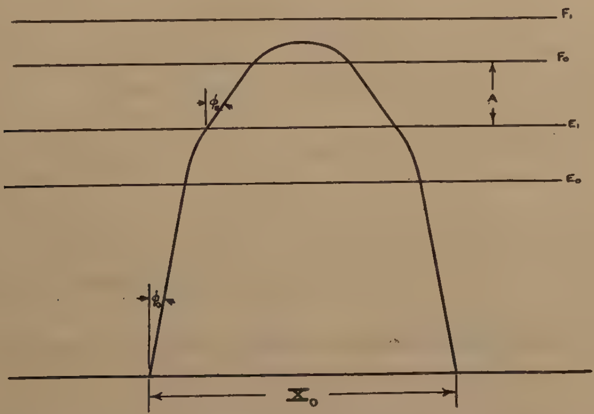

Fig. 12—Form of trajectory of ray in presence of two layers.

The square of the index of refraction is then assumed to remain constant at the value m^2 until a height F_0 is reached, when it again starts to decrease linearly to a value zero at a height F_1.

The layers are suggested schematically in Fig. 11 where the equations corresponding to the assumed linear deviations are also indicated. A curve showing the trajectory and indicating the notation adopted is given in Fig. 12. In carrying out the computations of virtual height for this case, free use will be made of the notation and the results evolved in a previous paper.[11]

In computing the virtual height h' we shall avail outselves of Breit and Tuve's theorem (1) which tells us that in the absence of dissipation (for any assumed variation of index of refraction with height), the group retardation is the same as that encountered by a ray traveling with the velocity of light, and traversing a triangular path with the

[10] G. Breit, "Group velocity and retardation of echoes," Proc. I.R.E., 17, 1508–1512; September, 1929.

[11] *See* equation $V-1$, p. 720, Kenrick and Jen, "Measurement of the height of the Kennelly-Heaviside Layer," Proc. I.R.E., 17, 711–733; April, 1929.

given base and given angle of departure. Hence (see Fig. 12 for notation),

$$X_0 = 2h' \tan \phi_0 \tag{1}$$

This gives us a relation between the virtual height h', the angle of departure ϕ_0, and the base X_0.

We shall now compute the relation between the coördinate $X_0/2$ and ϕ_0 by adding together the x intervals corresponding to each part of the path. This will enable us to compute the relation between actual height of the strata, the virtual height, m, ϕ_0, etc.

The ray moves in a straight line as far as the height E_0; hence its horizontal displacement thus far is merely

$$(\Delta x)_{E_0} = E_0 \tan \phi_0 \tag{2}$$

The ray travels in a parabolic arc through the E layer. Its horizontal displacement during this part of its motion is[10]

$$(\Delta x)_{E_1-E_0} = \frac{2 \sin^2 \phi_0}{1 - m^2}(E_1 - E_0)\left\{\cot \phi_0 - \sqrt{\cot^2 \phi_0 - \frac{(1 - m^2)}{\sin^2 \phi_0}}\right\} \tag{3a}$$

or,

$$(\Delta x)_{E_1-E_0} = \frac{2 \sin \phi_0 \cos \phi_0}{1 - m^2}(E_1 - E_0)\left(1 - \sqrt{1 - \frac{1 - m^2}{\cos^2 \phi_0}}\right) \tag{3b}$$

The ray leaves the E layer at angle given by the Snell law and the differential equation of the ray path (see reference 11; equation III–1) that is,

$$\sin \phi_E = \frac{\sin \phi_0}{m} \tag{4a}$$

from (4a) we have, by elementary trigonometry, that

$$\cos \phi_E = \frac{\sqrt{m^2 - \sin^2 \phi_0}}{m} \tag{4b}$$

and hence,

$$\tan \phi_E = \frac{\sin \phi_0}{\sqrt{m^2 - \sin^2 \phi_0}}. \tag{4c}$$

Using (4a) and (4c), and denoting the distance between the layers $(F_0 - E_1)$ by A, we have immediately that the x displacement of the ray in passing from E_1 to F_0 is:

$$(\Delta x)_{F_0-E_1} = (F_0 - E_1) \tan \phi_E = \frac{(A) \sin \phi_0}{\sqrt{m^2 - \sin^2 \phi_0}}. \tag{5}$$

Noting from (4) that the thickness of penetration $(Y_f - F_0)$ into the F layer is until $n^2 = \sin^2 \phi_E$, application of equation ($V-$1a, refer-

[10] *Loc. cit.*

ence 11) to this case gives us (in a manner similar to that employed in deducing (3)) that for the F layer, the x displacement will be

$$(\Delta x)_F = \frac{2\sin^2\phi_E(F_1 - F_0)}{m^2}\left\{\cot\phi_E - \sqrt{\cot^2\phi_E - \frac{m^2 - \sin^2\phi_E}{\sin^2\phi_E}}\right\} \quad (6a)$$

or making use of equations (4) to eliminate ϕ_E

$$(\Delta x)_F = (2\sin^2\phi_0)\left(\frac{(F_1 - F_0)}{m^2}\right)\left\{\frac{\sqrt{m^2 - \sin^2\phi_0}}{\sin\phi_0} - \sqrt{\frac{m^2 - \sin^2\phi_0}{\sin^2\phi_0} - \frac{m^2 - \dfrac{\sin^2\phi_0}{m^2}}{\dfrac{\sin^2\phi_0}{m^2}}}\right\} \quad (6b)$$

or,

$$(\Delta x)_F = \frac{2\sin\phi_0(F_1 - F_0)(\sqrt{m^2 - \sin^2\phi_0})}{m^4}\left[1 - \sqrt{1 - m^2}\right] \quad (6c)$$

Adding together (2), (3b), (5), and (6) to obtain the total $x_0/2$ (and utilizing (1) involving the virtual height), we have:

$$\frac{x_0}{2} = h'\tan\phi_0 = E_0\tan\phi_0 + (2\sin\phi_0\cos\phi_0)\frac{E_1 - E_0}{1 - m^2}\left(1 - \sqrt{1 - \frac{1 - m^2}{\cos^2\phi_0}}\right)$$
$$+ A\frac{\sin\phi_0}{\sqrt{m^2 - \sin^2\phi_0}} + \frac{2\sin\phi_0(F_1 - F_0)\sqrt{m^2 - \sin^2\phi_0}}{m^4}\left[1 - \sqrt{1 - m^2}\right] \quad (7a)$$

Hence,

$$h' = E_0 + 2\cos^2\phi_0\left(1 - \sqrt{1 - \frac{1 - m^2}{\cos^2\phi_0}}\right)\frac{E_1 - E_0}{1 - m^2} + \frac{A\cos\phi_0}{\sqrt{m^2 - \sin^2\phi_0}}$$
$$+ \frac{2(F_1 - F_0)\cos^2\phi_0\sqrt{m^2 - \sin^2\phi_0}\left[1 - \sqrt{1 - m^2}\right]}{m^4}. \quad (8)$$

So far, no approximations have been made, and (8) may be used to solve for h' for any x (by successive approximations).

However, when X_0 is small, that is, when we are working at nearly normal incidence, and when m^2 is small with respect to unity but large with respect to $\sin^2\phi_0$, equation (8) may be written approximately as

$$h' = E_0 + \frac{2(E_1 - E_0)}{1 + m} + \frac{A}{m} + \frac{F_1 - F_0}{m}. \quad (9a)$$

If now we define the critical frequency as that frequency for which the index of refraction of the E layer is zero, then for slightly higher frequencies, m will be small and the effective distance between layers may be increased to several times its actual value as may be seen from (9a). This effect, then, offers a possible explanation for the erratic heights observed for the F layer.

The relation between m and frequency may be written in a convenient form for nondissipative media (which, of course, are assumed throughout). Thus, the refractive index n is given by the well-known equation:[9]

$$n^2 = \epsilon = 1 - \frac{Ne^2}{m_1 \pi f^2} \tag{10}$$

where,

f = frequency,
ϵ = effective dielectric constant,
m_1 = electron mass,
e = electron charge,
N = electron density in electrons/cc

Thus from our definition and by (10) we get the following expression involving the critical frequency:

$$n^2 \equiv 0 = 1 - \frac{N_{(\max)E} e^2}{m_1 \pi f_c^2} \tag{11a}$$

or,

$$f_c^2 = \frac{N_{(\max)E} e^2}{m_1 \pi} \tag{11b}$$

where $N_{(\max)E}$ is the maximum electron density of the E layer. f_c is the highest frequency for which energy is returned by refraction at normal incidence from a nonturbulent layer of the type assumed in this analytical investigation.

On substituting f_c^2 for $(N_{(\max)E}\, e^2/m_1\pi)$ in (10) the square of the index of refraction at the highest point of the E layer is given by:

$$m^2 \equiv n^2_{(\min)E} \equiv 1 - \left(\frac{f_c}{f}\right)^2 \tag{11c}$$

If for a particular time we take $f_c = 3750$ kc, then at 4000 kc $m = \sqrt{1-(3.75/4)^2} = 0.35$ and the distance between layers indicated by height measurements would be about three times too great. Thus this critical value indicates an explanation for the virtual heights of 500 km or more, which were frequently observed during the tests. The critical frequency, which is proportional to the square root of the electron density, is extremely variable and doubtless changes considerably from hour to hour.

[9] *Loc. cit.*

The theory developed above would permit the coexistence of E and F layer reflections only very near the critical frequency. Thus just below the critical frequency no energy will get through the E layer for a given angle of incidence and a reflection from this layer will occur.

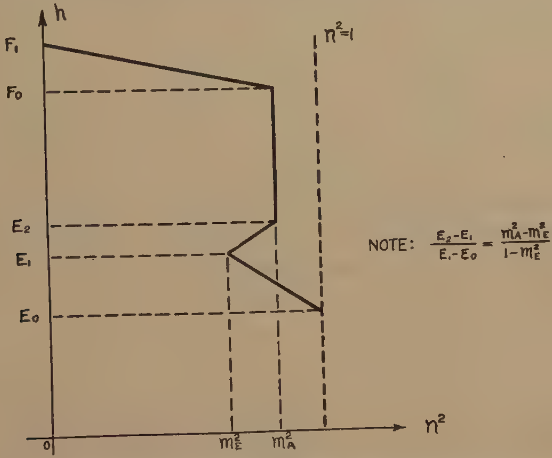

Fig. 13—More general form of assumed n variation showing minimum value in first layer.

However, at a slightly smaller angle of incidence energy can get through the E layer and be reflected from the F layer. During the tests described such coexisting reflections were observed at various

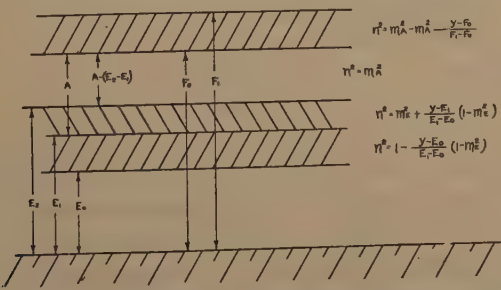

Fig. 14—Layer distribution corresponding to more general variation of n indicated in Fig. 13.

times on 3000, 4100, and 5000 kc so that the value of f_c chosen may be considered as a fair example for calculating the maximum electron density in the E layer by means of (11b). This calculation gives:

$$N_{\text{MAX}}\ (E\text{ layer}) = 1.7\times10^5 \text{ electrons per cc.}$$

A similar computation for the summer F layer, assuming value $f_c=5500$ kc gives:

$$N_{\text{MAX}}(F\text{ layer}) = 3.8\times10^5 \text{ electrons/cc.}$$

Little specific information is available as to the electron distribution in the upper atmosphere, but it is perhaps appropriate to consider a slight generalization of the above analysis, for, thus far, it will be noted we have considered the reduction of n^2 with height to be progressive; that is, no maximum and subsequent minimum of N was considered between the layers. A more general type of variation to which the above analysis may be readily extended is obtained by an assumption such as indicated in Fig. 13.

This variation differs from that of Fig. 10 only in that the n^2 variation is assumed to reach a minimum m_E^2 in the E layer and then n^2 is again assumed to increase (along a symmetrical straight line of equal slope) to a value m_A^2, where it remains until the height F_0 is attained. The equation of the variations is indicated in Fig. 14, and the generalization of (7) to include this case gives:

$$
\begin{aligned}
\frac{x_0}{2} &= h' \tan\phi_0 = E_0 \tan\phi_0 \\
&+ 2\left[2 \sin\phi_0 \cos\phi_0 \frac{E_1 - E_0}{1 - m_E^2}\left(1 - \sqrt{1 - \frac{1 - m_E^2}{\cos^2\phi_0}}\right)\right. \\
&\quad - 2 \sin\phi_0 \cos\phi_0 \frac{(E_1 - E_0)}{1 - m_E^2}\left(1 - \sqrt{1 - \frac{1 - m_A^2}{\cos^2\phi_0}}\right) \\
&\quad + \frac{[A - (E_2 - E_1)] \sin\phi_0}{\sqrt{m_A^2 - \sin^2\phi_0}} \\
&\quad + \frac{2 \sin\phi_0 (F - F_0)\sqrt{m_A^2 - \sin^2\phi_0}}{m_A^4}\left[1 - \sqrt{1 - m_A^2}\right].
\end{aligned}
$$

In writing this equation we note that, for this case, by symmetry, the parabolic path $(\Delta x)_{E_0-E_2}$ would be just twice its value in (7a) *provided* $m_A^2 = 1$. For general variations corresponding to other values of m_A^2, $(\Delta x)_{E_0-E_2}$ is twice that of (7a) minus the (Δx) corresponding to the change from $n^2 = 1$ to $n^2 = m_A^2$ along the parabolic path; this is represented by the subtractive term of (7b).

The approximations previously employed in deducing (9) from (7) may be applied (subject to the same approximations) to deduce from (13) the value of virtual heights h' for the generalized case, that is,

$$
h' = E_0 + \frac{2(E_1 - E_0)}{1 - m_E^2}(1 - 2m_E + m_A) + \frac{A - (E_2 - E_1)}{m_A} + \frac{F_1 - F_0}{m_A} \qquad \text{(9b)}
$$

It will be noted that when $m_A = m_E$ and $(E_2 - E_1) = 0$, equations (7b) and (9b) reduce respectively to (7a) and (9a).

The nature of the phenomena indicated by (7b) and (9b) is not essentially different from those previously discussed but indicate that for a given $F_1 - E_0$ and m_E the variation in virtual height of the F layer produced by the E layer is greatest when the maximum m_A^2 between the layers is small (that is, when there is no pronounced decrease in N between the layers). The values indicated in the discussion of (9) are thus upper bounds (at least for the types of variations assumed). Previous investigations have indicated the conclusions to be independent of the exact form of variations assumed,[4] except for certain very special conditions,[10] and the relations deduced above may hence be considered as reasonably representative. In general, we should expect the value of m_A to be distinctly less than unity, so the phenomenon discussed will be in evidence, in general, near the critical frequency for the E layer. (Note that if m_A is *not* considerably less than unity, (7b) should be used, as (9b) is no longer a good approximation.)

2. Reflection Phenomena

The consideration of refraction in the presence of two layers, as outlined above, throws considerable light on some of the phenomena observed in Kennelly-Heaviside layer observations. The presence of reflections simultaneously from both layers may be explained as above for frequencies very near the critical frequency or they may be explained by postulating a nonuniform distribution; (that is, turbulent clouds of electrons with greater density than the average) which return some of the energy while allowing the remainder to travel to the F layer. Phenomena of this sort doubtless are of importance but seem hardly adequate to explain all the results. Numerous cases of the type shown in Fig. 1b may be found, where rays apparently returned from the E and F regions, coexist.

An investigation of reflection coefficients seems appropriate in this connection. Numerous investigators have emphasized that little reflection will occur at gradually varying boundaries provided the first and second derivatives of the variation are not large and do not exhibit discontinuities, in short, provided the variation extends over a number of wavelengths and is free from abrupt stratifications.[9,12,13]

It must be remembered, however, that the power returned is usually small and so variable that reflection coefficients of a few per cent would be sufficient to explain many of the observed phenomena. It

[4] *Loc. cit.*
[10] *Loc. cit.*
[12] Rayleigh (Strutt, J. W.) "On reflection of vibrations at the confines of two media between which the transition is gradual," *Proc. London Math Soc.*, pp. 51–56; 1881.
[13] Rayleigh (Strutt, J. W.), "On the propagation of waves through a stratified medium with special reference to the question of reflection." Collected Works, VI, pp. 71–90.

therefore seemed appropriate to review quantitatively the reflection theory with a view to establishing how much this phenomenon might be invoked to explain the results. In this paper we will confine ourselves for simplicity to the case corresponding to most of the observations; that is, that of normal incidence.

Lord Rayleigh's work is notable among the investigations of this problem of the reflection at normal incidence from gradually varying media. He obtained solutions directly from the differential equation of the motion (for certain special assumed laws of variations of the medium which rendered the resulting equation integrable).[12] He also

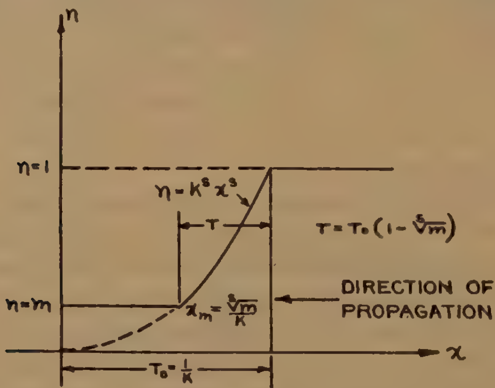

Fig. 15—Form of transition assumed in investigation of reflection (Case I).

obtained the solution as a limiting case of the solution for stratified transitions as the number of stratifications became infinite and the amplitude of the transitions between any two strata infinitesimal.[13]

The most general (and for our purposes, the most convenient) relation he derives is to be found in the second paper.[13] Rayleigh's investigations are not limited to waves in optical media but, as he points out, may be readily reduced to this case. When so reduced, this equation may be written in our notation as:

$$\frac{B_1}{A_1} = -\int \frac{dn}{2n} \epsilon^{-2j\int_0^x (2\pi n/\lambda n_0)\,dx} \tag{12}$$

where,

n = index of refraction at any point x in the variable medium.
λ = wavelength *in vacuo*.
$j = \sqrt{-1}$
B_1 = amplitude of reflected wave.
A_1 = amplitude of incident wave.
r = reflection coefficient at normal incidence.

[13] *Loc. cit.*
[13] *Loc. cit.*, page 80, equation 5.

It should be noted that $|B_1/A_1| = r$. This relation will be used in this investigation.

The results, of course, depend on the assumed type of variation of dielectric constant, but an idea of the magnitude to be expected may be obtained by investigating some plausible distributions.

We shall now proceed to evaluate (12) for rather general types of variation of n with x.

Case I

We shall first consider a variation of the form $n = k^s x^s$ (see Fig. 15).

Substituting this value of n in (12) we have for the reflection (for a strata varying from an $n=1$ to an $n=m$ along the assumed curve)

$$\frac{B_1}{A_1} = - \int_{x=1/k}^{x=\sqrt[s]{m}/k} \frac{s\,dx}{2x} \epsilon^{j(4\pi/\lambda)[(k^s x^{s+1}/s+1)-(k^s/s+1)]} \tag{13}$$

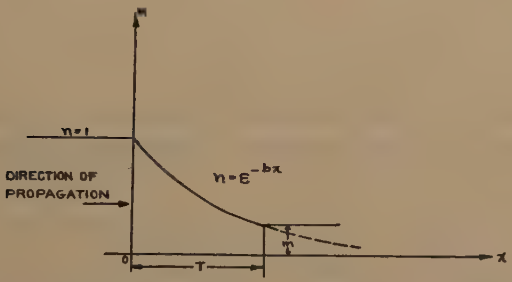

Fig. 16—Form of transition assumed in investigation of reflection (Case II).

We may change the variable by letting $x^{s+1} = z$. Making this change and dealing only with the magnitude of B_1/A_1; (that is, neglecting their relative phase) we have, considering only magnitudes

$$r = \left|\frac{B_1}{A_1}\right| = \left|\frac{s}{2(s+1)} \int_{z=1/k^{s+1}}^{z=m^{(s+1)/s}/k^{s+1}} \frac{dz}{z} \epsilon^{j(4\pi/\lambda)[(k^s z)/(s+1)]}\right| \tag{14}$$

We are thus only required to evaluate an integral of the relatively simple form

$$\int \frac{\epsilon^{ay}}{y}\,dy \tag{15a}$$

Employing successive integration by parts we may write:

$$\int \frac{\epsilon^{ay}}{y}\,dy = \frac{\epsilon^{ay}}{ay} + \frac{\epsilon^{ay}}{a^2y^2} + 2\int \frac{\epsilon^{ay}}{a^2y^3}\,dy. \tag{15b}$$

For real values of y and pure imaginary values of a this series may be written in the asymptotic form:

$$\left|\int \frac{\epsilon^{ay}}{y}dy\right| = \left|\frac{\epsilon^{ay}}{ay} + \frac{\epsilon^{ay}}{a^2y^2}\right| + 2R\int\frac{dy}{a^2y^3} \tag{15c}$$

$$-1 \leqq R \leqq 1.$$

Remembering that a is a pure imaginary, we note that the integral in general represents a rapidly pulsating function but that always

$$\left|\int \frac{\epsilon^{ay}}{y}dy\right| \leq \left|\frac{1}{ay}\right| + \left|\frac{1}{a^2y^2}\right| + \left|\frac{1}{a^2y^2}\right| \tag{15d}$$

Applying (15d) to integral 14, and further approximating by neglecting phase differences at the limits, we may write further the order of magnitude of (B_1/A_1) by the inequality

$$r = \left|\frac{B_1}{A_1}\right| \leqq \frac{s}{2(s+1)}\left\{\left|\frac{(s+1)k\lambda}{4\pi}\left(\frac{1}{m^{s+1/s}}+1\right)\right.\right.$$
$$\left.\left.+ 2\frac{(s+1)^2k^2\lambda^2}{16\pi^2}\left(\frac{1}{m^{2(s+1)/s}}+1\right)\right|\right\}. \tag{16}$$

For $1/m$ large compared with unity and $(s+1)\lambda k/4\pi m^{s+1/s}$ small compared with unity, this inequality approaches an equality to terms of the order of r^2, the function ceases to oscillate violently, and we may write (noting further that $T/T_0=(1-m^{1/s})$,

$$r \doteq \frac{sk\lambda}{8\pi m^{s+1/s}} = \frac{s\lambda}{8\pi m^{s+1/s}T_0} = \frac{s\lambda}{8\pi T}\ \frac{(1-m^{1/s})}{m^{s+1/s}} \tag{17}$$

where,

s = exponent of law of variation
λ = wavelength (km)
T = thickness of varying layer (km)
m = minimum refractive index n encountered during variation in layer.

This coefficient does decrease with λ/T_0; that is, with the ratio of the wavelength to the thickness of the layer (extrapolated to zero n), but it will be noted the decrease is only linear; that is, for small coefficients of reflection the coefficient r is inversely proportional to the first power of the thickness of the layer. This coefficient varies (but not critically) with the powers of the variation of n with x. Thus, for a linear variation $(s=1)$ we have, taking $T=10$ km, $\lambda=0.1$ km, and $m=0.3$; $s(1-m)/8\pi(m^2)=0.3$ and $r=0.003$. For $s=2$ the other data remaining the same, we have $s(1-m^{1/2})/8\pi(m^{1/2})=0.2$ and $r=0.002$; likewise for $s=3$, $r=0.002$.

While these coefficients of reflection are small, they are nevertheless quite appreciable, and reflections thus produced should be quite observable. It will be noticed further that close to the critical frequency much larger values of r may be found due to the small values of n en-

countered in that region. (However, note that the equation for r is of course not valid unless $r^2 \ll r$).

Case II: $n = \epsilon^{-bx}$

We shall next consider an assumed variation of the form $n = \epsilon^{-bx}$.

As will be noted by reference to Fig. 16, we shall consider an exponential variation starting at $x=0$ and extending to $x=T$, where the minimum index of refraction ($n=m$) is encountered.

This gives $\epsilon^{-bT} = m$ or $b = \log_E(1/m)/T$.

Equation (12) when applied to this case gives:

$$r = \left|\frac{B_1}{A_1}\right| = \left|\frac{b}{2}\int_{x=0}^{x=T} \epsilon^{j(4\pi/\lambda b)} dx^{\epsilon^{-bx}}\right| \tag{18}$$

or, letting $\epsilon^{-bx} = y$

$$r = \left|\frac{1}{2}\int_{y=1}^{y=m} \frac{dy}{y}\epsilon^{j4\pi y/\lambda b}\right|. \tag{19}$$

This integral is identical in form to (15) and the same methods of solution hold; hence by (15b) and approximations similar to those employed in case I we have for r small and $m \ll 1$.

$$r = \left|\frac{B_1}{A_1}\right| = \frac{\lambda \log_\epsilon m}{8\pi T m}. \tag{20}$$

which holds to terms of the order of r^2.

Taking $\lambda = 0.1$ km, $m = 0.3$, $T = 10$ as before, (20) gives $r = 0.001$, a result in good accord with that previously obtained.

Case III: $n = P/X$

It is of interest, in conclusion, to compare the above solutions with that obtained by Rayleigh in his earlier paper.[14] This solution is of particular interest because it was not obtained by use of (12) but by a direct investigation of the wave equation, which is solvable in this special case. (It will be noted that $s = -1$ is excluded from our sth power solution and has to be handled separately due to the logarithm obtained on integrating dx/x.) Comparison with our results obtained above requires only a reduction of his general solution for any motion satisfying the wave equation to our special notation.

Rayleigh's problem, when reduced to the optical case and our notation, is indicated in Fig. 17. Propagation is assumed in a positive direction and the variability to extend between $x = x_1$ and $x = x_2$ (we have

[14] *Loc. cit.*, p. 55.

taken $x_1=1$ and hence $x_2=1/m$ which, if T is the desired layer thickness in km, necessitates a unit of length of $T/[(1/m)-1]$ km.)

In these units λ is related to T in km by $\lambda=(\lambda\text{ km})\,([(1/m)-1]/T)$. In examining units in the solution as given by Rayleigh, care must be taken not to confuse our m (minimum value of n) with the m he uses, which we shall designate by m' to avoid confusion.

$$\left(m' = \sqrt{\left[\frac{2\pi T}{\lambda_{km}\left(\dfrac{1}{m}-1\right)}\right]^2 - \frac{1}{4}}\right)$$

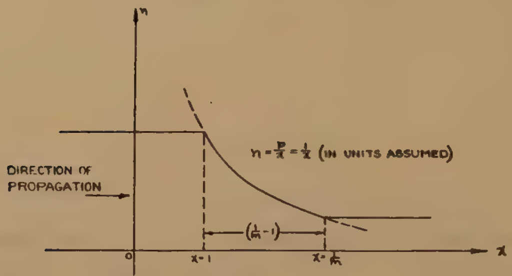

Fig. 17—Form of transition assumed in investigation of reflection (Case III).

In our notation, Rayleigh's solution is

$$r = \left|\frac{B_1}{A_1}\right| = \sqrt{\frac{\sin^2\left(m' \log \dfrac{1}{m}\right)}{4m'^2 + \sin^2\left(m' \log \dfrac{1}{m}\right)}}. \tag{21}$$

It is interesting to note that this solution pulsates rapidly from zero to maxima of the order of:

$$r \doteq \frac{\lambda_{km}}{4\pi T_{km}\, m} \tag{22}$$

$(m \ll 1)$ and $(r \ll 1)$.

This upper bound is a result of the same order of magnitude as obtained in cases I and II. An application of the integral given in (12) will indicate this property, for in this case changes of variable reduce it to a form involving an integral between limits of the form $K\left|\int_A^x \epsilon^{ik\theta} d\theta\right|$ which does not involve an inverse power of θ multiplying the exponential, and hence continues to pulsate reading zero at frequent

intervals over all finite but large ranges of x. Space will not be taken here for a complete discussion of this case, as the result is obtained by Rayleigh by the independent method which offers a further check on the order of magnitude of our results obtained in the other cases. Thus (22) gives for $\lambda=0.1$ km, $T=10$ km, and $m=0.3$ an upper bound of r of

$$r = 0.003.$$

VI. Conclusions

The results of this paper indicate that the phenomena observed may be profitably investigated both from the point of view of refraction in the presence of two layers and from the point of view of reflection (particularly near the critical frequency f_c where the minimum value of the effective index of refraction nearly attains zero). Theoretical predictions that the most marked effects due to these causes are to be noted near the critical regions, are apparently justified by the experiments reported, which also seem to confirm the presence of two rather well-defined layers or regions of ionization. The development of the theory has given an indication of what may be looked for and further experiments should be conducted with these things in mind.

While the reflection coefficients computed are, in fact, small (except quite near the critical frequency), and vary somewhat according to the law of variation of n assumed, they apparently may reach several per cent on the basis of plausible assumption as to laws of variation of n, thickness of layer, wavelength, etc., and it hence seems appropriate that this effect should be borne in mind as a possible explanation of some of the phenomena observed for, except very near the critical frequency, refraction can not explain the simultaneous appearance of rays from both layers unless horizontal gradients; that is, electron clouds, are assumed. While such clouds are, of course, possible, it must be borne in mind that they would not be stable and hence would be expected to exist only under disturbed conditions.

With present high sensitivity radio receivers, an attenuation ratio of 100:1 corresponding to a reflection coefficient of 0.01 should, under favorable conditions, be adequate provided the general signal-to-noise level were sufficiently high (that is, high power transmission under favorable conditions). Diurnal changes in sky wave energy of at least this order of magnitude are frequently observed on many frequencies.

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FREQUENCY STABILIZATION OF RADIO TRANSMITTERS*

By

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Summary—*The importance of maintaining constant the frequency of radio transmitters has been universally recognized and the investigation of methods of frequency stabilization was left as a pending question after the first meeting of the C.C.I.R.*

The quartz-control system with thermostatic temperature control of the quartz crystal should be utilized when the highest precision (of the order of 0.001 per cent) is required in the constancy of the transmitted frequency. However, the system necessitates a complicated and expensive installation and at the same time the frequency of transmitters is not so flexible. When the highest precision is not of primary importance the following methods devised by the writers may be applied with the advantages of simplicity and lower cost of installation. These methods, fully described in the present paper, may readily be applied to a master oscillator so as to give a frequency stability of the order of 0.01 per cent.

(a) *Constant-frequency oscillator.*

In an ordinary self-oscillator, the effect of the variation in supply voltages may be minimized by simple means. Two methods have been developed for the purpose:

1. *Resistance-stabilized oscillator.*
2. *Phase-compensated oscillator.*

(*b*) *Quartz-stabilized oscillator.*

A quartz crystal is introduced in the grid circuit of an oscillator in such a manner that it is very loosely coupled to the oscillator but still retains its stabilizing function. Power output of the oscillator can be obtained up to 100 watts or more and thus the number of stages of amplifiers to be used in a transmitter can be reduced.

(*c*) *Mechanically-stabilized oscillator.*

A vernier condenser, mechanically driven and controlled by a relay, is connected to the oscillator circuit. A quartz oscillator is used as the frequency standard, and the relay is actuated by beat frequency and controls the rotation of the vernier condenser so as to counteract the frequency variation. Thus a high power oscillator can be directly stabilized by a low power quartz-oscillator.

(*d*) *Valve-stabilized oscillator.*

A valve is coupled to the oscillator in such a manner that it acts as a pure capacity, and the effective capacity is controlled by the grid-bias voltage. A quartz resonator is used as the frequency standard and its resonance characteristic is utilized to control the grid-bias voltage of the above-mentioned valve so as to stabilize the oscillator frequency. This method is also applicable to a high power master oscillator.

I. Introduction

THE importance of maintaining constancy in the frequency of a transmitter has recently been considered very seriously, since the number of radio stations has considerably increased due to the advent of a variety of new services in radio communication.

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In order to cope with the situation, permissible tolerances in holding the transmitters at their assigned frequencies have been given and "the study and perfecting of methods technically available for maintaining constant the frequency of a transmitter" was left as a pending question at the first meeting of the C. C. I. R. held at the Hague in 1929.

For stabilizing the frequency of a transmitter, there have been devised a number of schemes, the most widely-used ones being the quartz-control system and the valve master-oscillator system.

The quartz-control system has been universally recognized as the best means for the accurate maintainance of frequency at an assigned value, and it is unquestionably the only method to realize a constancy of the order of 0.001 per cent at the present state of the technique. But it has the drawback of requiring a complicated installation and accordingly a very expensive transmitter, because the power output of a quartz oscillator is limited to a certain low value and a multistage power amplifier is needed, and moreover it has no flexibility in the frequency of transmission.

The valve master oscillator, on the other hand, has the merit of flexibility of frequency and also that of simplicity of installation, if it is operated at a relatively large power output, but its most serious demerit lies in the fact that the frequency is subject to fluctuation and drift due to variations in the supply voltages and changes in the circuit constants.

Thus it may be concluded that when a very high degree of precision is required in maintaining the frequency constant, the quartz-control system with thermostatic temperature control of the quartz crystal should be recommended, but for the more general usage not requiring such a high degree of precision, the valve master-oscillator system of high power output may be adopted with advantage, especially from the point of view of cost, if some simple means can only be provided with which to stabilize the frequency within a permissible limit.

The writers investigated the problem by carrying out considerable experiments and obtained the satisfactory methods reported in the present paper.

II. Constant-Frequency Oscillator

Frequency variation in valve oscillators may have two different characteristics according to its main cause, the first due to changes in the external circuit constants and the second due to changes in the internal constants of the valve itself.

stitute, June 24, 1931. Presented before U.R.S.I., Copenhagen, 1931. Printed by National Research Council of Japan.

The change in external circuit constants is due to mechanical vibration or shock, a change of temperature, humidity or atmospheric pressure, etc., so that a frequency change due to this cause is gradual in character as long as mechanical rigidity is secured and other variable elements such as the antenna are not involved in the oscillator circuit.

The second cause, a change of valve constants, is mainly due to a change of internal plate resistance resulting from changes in plate or

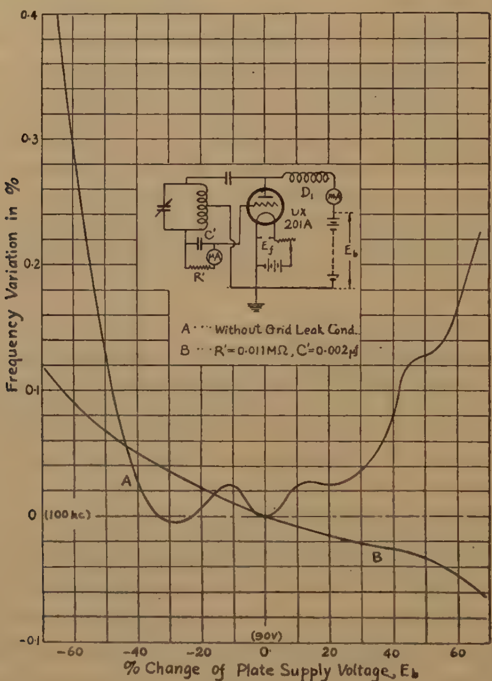

Fig. 1A.

filament supply voltage or static grid bias, and the variation is often instantaneous if the supply sources are not well regulated.

One of the writers carried out an experimental study on the frequency-variation characteristics of the second kind using receiving valves, and obtained two schemes of frequency stabilization, very simple in installation and readily applicable to existing oscillators.

(*a*) *Resistance-Stabilized Oscillator*

The frequency of a valve oscillator is in the main determined by the natural frequency of the oscillation circuit, but is a little affected by variations in plate and filament voltages. Although the percentage

variation of frequency due to this cause is small compared to the absolute value, it is still responsible for unreliable service and interference to other stations if an ordinary self-oscillator is used in a radio transmitter without paying particular attention to the stabilization of frequency.

The principal cause of frequency variation due to supply voltages is the variation of the internal plate resistance of the valve, and there

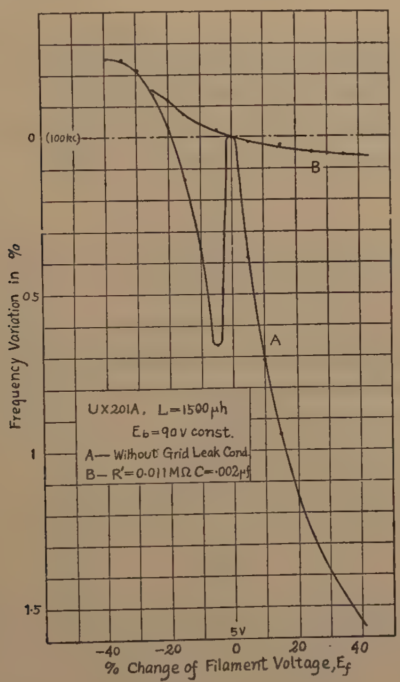

Fig. 1B.

have been devised some simple methods to minimize this effect. One of the widely-known methods is to introduce an external high resistance in the plate circuit in series with the oscillatory circuit, so that the oscillation frequency becomes less affected by the variation of the internal resistance.

Another method of minimizing the frequency variation is to utilize a grid leak of high resistance value for giving dynamic bias. The grid-bias voltage is then dependent on the intensity of oscillation and it tends to reduce automatically the change in the plate resistance caused by the variation of supply voltages (Fig. 1-A, Fig. 1-B and Table I).

TABLE I

Relation Between Frequency Variation and Grid Leak Resistance (R') in Hartley Oscillator; Valve UX-201A, Grid Condenser $C' = 0.00212 \mu f$.

Change of Supply Voltage	Frequency increase at 100 kc in 10^{-5}				
	$R'=0$	$R'=11{,}000\,\Omega$	$R'=20{,}620\,\Omega$	$R'=81{,}840\,\Omega$	$R'=165{,}300\,\Omega$
10 per cent decrease of plate voltage, E_b ($E_f=5$v const.)	140	10	8	4	3.5
10 per cent decrease of filament voltage, E_f ($E_b=90$v const.)	500	40	32	8	2.5

These are the simple and effective methods of reducing frequency variation which have already been revealed by many authors, but they are not capable of producing complete stability, no matter how the resistance in the plate circuit or the grid leak is increased, and the high external plate resistance affects the power output.

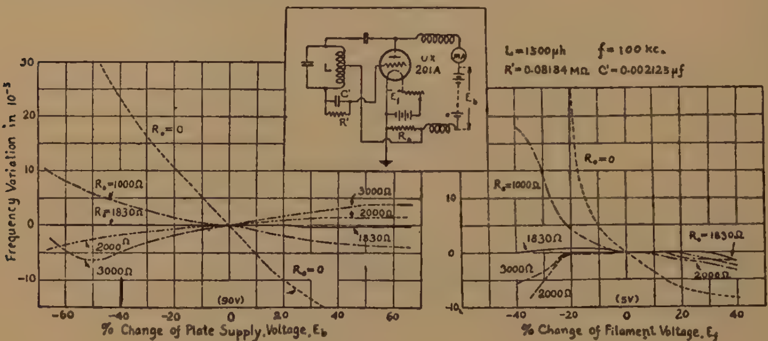

Fig. 2.

The circuit devised by one of the writers, which may be called the "resistance-stabilized oscillator," is shown in Fig. 2. The external plate resistance R_0 is introduced in such a way that it not only acts as a resistance in the oscillating plate circuit, but produces at the same time a part of the grid bias voltage and of grid excitation on account of the potential drop produced by the direct current and oscillating components of the plate current flowing in it. It may be looked upon that the stabilizing action of the external resistance and that of the grid dynamic bias are combined to give stability higher than that obtainable by making use of only one of them.

The curves in Fig. 2 show the frequency variation characteristics of an oscillator with respect to different values of the stabilizing resistance R_0, obtained with a receiving valve UX–201-A at the frequency of 100 kc using a suitable value of grid leak, and show that an optimum value of R_0 is in this particular case 1830 ohms. The variation characteristic

is reversed by further increase of R_0, and thus complete stability is obtainable at this optimum value of the resistance. In the above example the order of stability is within 0.001 per cent for a very wide variation in supply voltages.

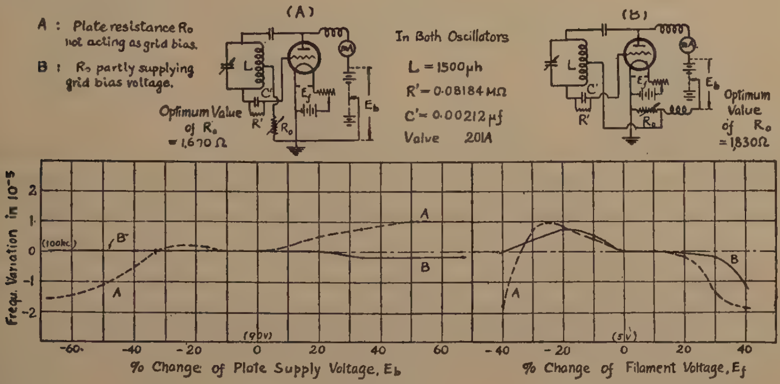

Fig. 3.

If the external plate resistance has no additional action of automatic bias control, the oscillation frequency becomes less stable as shown in curve A in Fig. 3 and moreover no reversing effect is observed.

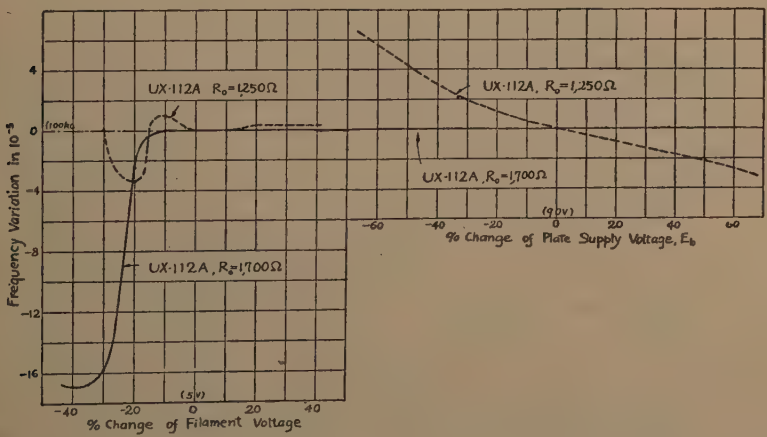

Fig. 4.

The value of stabilizing resistance giving the minimum frequency variation differs slightly in accordance with the type of the tube, even though the same oscillation circuit is used, as shown in Table II.

It should be noted that the optimum tube of the external plate resistance R_0 for the plate voltage variation does not always coincide

TABLE II

OPTIMUM VALUES OF THE RESISTANCE R_0 FOR DIFFERENT TYPES OF VACUUM TUBES, UNDER Z_p = 107,500 OHMS AND R' = 81,800 OHMS.

Vacuum tube			Optimum value of external plate resistance R_0 ohms
Type	Amplification constant	Internal resistance, ohms	
Cymotron UX-240	28.5	53,000	1,500
" UX-201-A	8.2	8,000	1,830
" UX-112-A	7.0	3,900	1,250
" UX-171-A	3.0	1,400	2,000

with the optimum value for the filament voltage variation. An example is shown in Fig. 4 using UX–112-A; for the plate voltage variation the best result is given at $R_0 = 1700$ ohms, but for the filament voltage variation a little better result is obtained at $R_0 = 1250$ ohms.

Although a portion of the grid bias voltage is due to the insertion of the resistance R_0, it only assists the main bias voltage produced by the grid leak in keeping the internal plate resistance in a limited range, and the use of a relatively high value of grid leak is essential for taking full advantage of the present method. As shown in Table III, with the grid leak $R' = 81{,}840$ ohms, the optimum value of R_0 is 2000 ohms; but with $R' = 10{,}430$ ohms, optimum R_0 giving such a stability cannot be found.

TABLE III

R' = 81,840 ohms, UX-171-A, L_p = 1500 μh, C' = 0.002125 μf

Filament		Frequency variation at 100 kc in 10^{-5}						
Volts	Change in per cent	$R_0=0$	$R_0=200$	$R_0=1000$	$R_0=1500$	$R_0=2000$	$R_0=2500$	$R_0=3000$
3.0	−40	44	14.2	3.75	2.6	1.7	1.7	1.7
5.0	0	0	0	0	0	0	0	0
7.0	+40	2.6	2.6	1.0	0	−0.75	−1.4	−2.1

R' = 10,430 ohms, UX-171-A, L_p = 1500 μh, C' = 0.002125 μf

Filament		Frequency variation at 100 kc in 10^{-5}			
Volts	Change in per cent	$R_0=1000$	$R_0=2000$	$R_0=5000$	$R_0=7000$
3.0	−40	59.95	22.15	4.9	4.25
5.0	0	0	0	0	0
7.0	+40	3.3	1.4	−0.2	−1.4

In a Hartley oscillator, a part of the oscillation circuit is shunted through the internal plate resistance. Consequently the effect of the variation of internal plate resistance on oscillation frequency becomes less if a low value of the plate-filament impedance of the oscillation circuit is selected, that is, if a smaller inductance and a larger capacity

are used. This fact is also indicated by α and β in formulas (1) and (2), respectively; and in such a case a lower value of stabilizing resistance can be used. These relations were obtained as shown in Table IV.

TALBE IV

RELATION BETWEEN Z_p AND OPTIMUM R_0.

Oscillation circuit		Plate impedance* Z_p ohms	Optimum value of inserted plate resistance R_0 ohms
Inductance L	Capacity C		
638 μh	0.00389 μf	36,300	1000
1500	0.00161	107,500	1830
2200	0.00113	156,800	2250

$$* \; Z_p = \frac{L}{CR}\left(\frac{L_p + M}{L}\right)^2, \qquad L = L_p + L_G + 2M.$$

The writers applied the present method to a power valve UV–204-A (rated plate voltage 2000 volts, filament 11 volts, output 250 watts), and obtained the result shown in Fig. 5, from which the optimum resistance R_0 is known to lie between 65 and 120 ohms. Another case using a power valve VM–1101 (rated plate voltage 10,000 volts, filament 16.5 volts, output 800 watts), gave similar results at $R_0 = 600$ ohms.

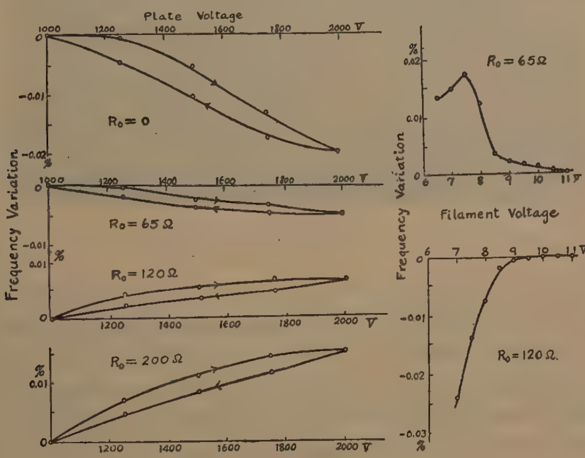

Fig. 5.

When power valves were used a frequency drift of a sluggish character was invariably observed, and in one example using a UV–204-A valve oscillating at 3000 kc, the frequency decreased continually from the time of starting. It took about an hour for the frequency to reach a stationary state and the total frequency drift was 0.05 per cent. When

the oscillator is operating at a lower frequency this frequency drift is principally caused by the variation of circuit constants due to heating, but in the above-cited case, especially at higher frequencies, the main cause was found to be the variation of the interelectrode capacity resulting from the heating of the valve. In order to avoid this sort of frequency drift, it is recommended that the valve be operated at light load and utilized after a stationary state in the temperature of the valve is reached.

The present stabilizing action was proved to be equally effective for the instantaneous variation of supply voltages and the oscillator could

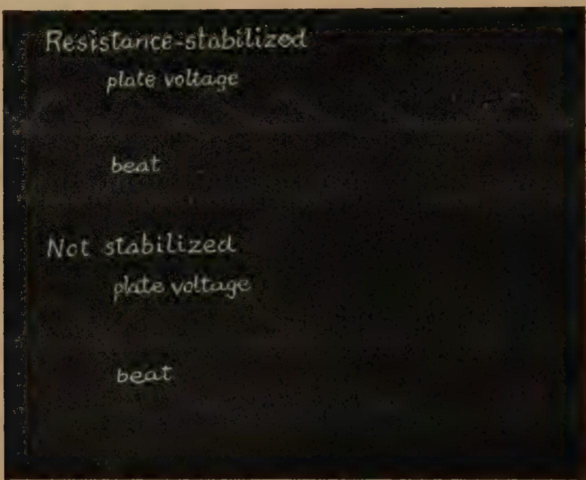

Fig. 6.

be subjected to modulation without varying in accordance with the frequency of modulation. The oscillogram shown in Fig. 6 is one of the results proving this fact. The present oscillator will also be of special value in this respect when applied to a modulated master oscillator.

The effect of the change of internal plate resistance on the oscillation frequency is different in character according to whether the circuit is of the Hartley or of the Colpitts type.

One of the writers has derived theoretical relations for the oscillating frequency which, as a first approximation, may be expressed in the following way:

For a Hartley circuit,

$$\omega^2 = \frac{1}{C(L_p + L_g + 2M) + \dfrac{L_pL_g - M^2}{kr_pr_g}}$$

hence,

$$\omega = \omega_0 \left(1 - \tfrac{1}{2}\alpha\omega_0^2\right) \qquad (1)$$

$\alpha\, \omega_0^2 \ll 1$ being assumed,

where,

$$\alpha = \frac{L_p L_g - M^2}{k r_p r_g}.$$

For a Colpitts circuit,

$$\omega^2 = \frac{1}{L}\left(\frac{1}{C_p} + \frac{1}{C_g}\right) + \frac{k}{r_p r_g C_p C_g}$$

hence,

$$\omega = \omega_0\left(1 + \frac{1}{2}\,\frac{\beta}{\omega_0^2}\right) \qquad (2)$$

$\beta/\omega_0^2 \ll 1$ being assumed,

where,

$$\beta = \frac{k}{r_p r_g C_p C_g}.$$

In both cases,

ω_0 represents the resonant frequency of the oscillatory circuit

r_p and r_g represent the internal plate and grid resistances, respectively

$$k = 1 - \frac{\mu}{\nu}$$

where,

μ represents the amplification factor.

ν represents the reflex factor $\left[-\left(\dfrac{\partial E_p}{\partial E_g}\right)_{I_g} = \text{const.}\right]$

and,

L_p L_g, M, C_p, and C_g represents the circuit constants.

Thus, in a Hartley circuit the oscillation frequency is always lower than the resonance frequency of the oscillatory circuit, and since the deviation is inversely proportional to the internal resistances r_p and r_g, an increase in the supply voltges, which ordinarily results in a decrease of r_p and r_g, causes a decrease in the oscillation frequency.

In a Colpitts circuit the oscillation frequency is always higher than the resonance frequency, and since the deviation is inversely proportional to r_p and r_g, a rise in the supply voltages causes an increase in the oscillation frequency. Variations of supply voltages in the above two

cases may have reverse effects if the rise of voltages causes an increase in r_p and r_g, which occurs in a valve operating at low emission.

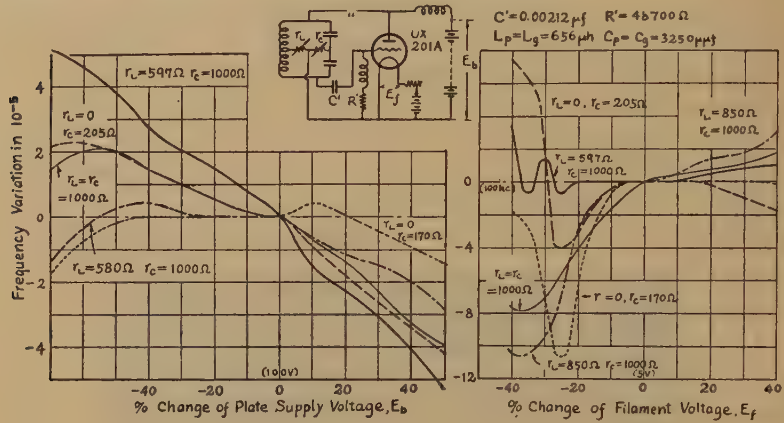

Fig. 7.

The cause of frequency variation, when considered physically, is attributed to the variation of the phase difference between plate and

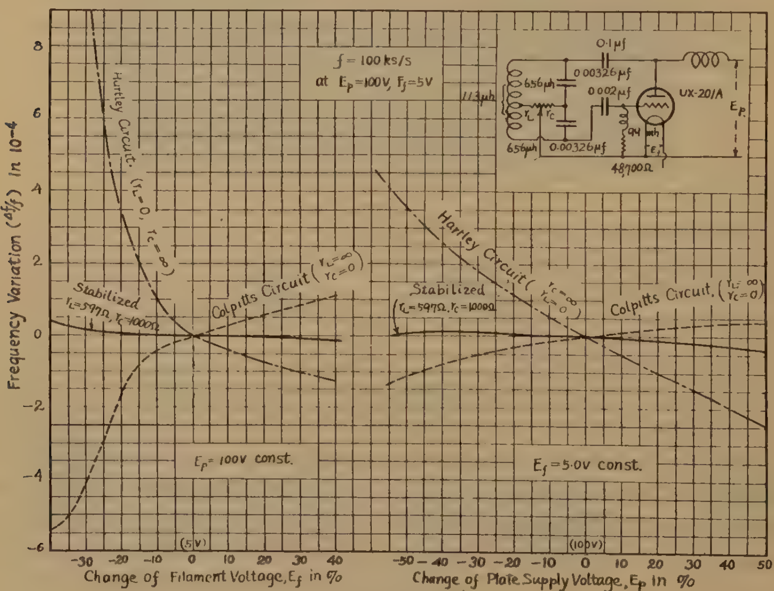

Fig. 8.

grid potentials due to the change of internal plate resistance, and in the two types of circuit mentioned above, phase shift in the grid potential

due to a change of r_p or r_g occurs in the opposite sense, hence resulting in their reversed frequency variation characteristics.

The present scheme of frequency-stabilization, which may be called the "phase-compensated oscillator," makes use of a means for controlling the phase of the grid potential with a circuit shown in Fig. 7. A resistance is connected across the oscillatory circuit in such a way as to connect equipotential points in the inductive and the capacitive legs of the oscillation circuit, and the filament is connected to a proper point on the resistance. Then it is possible to find a suitable proportion

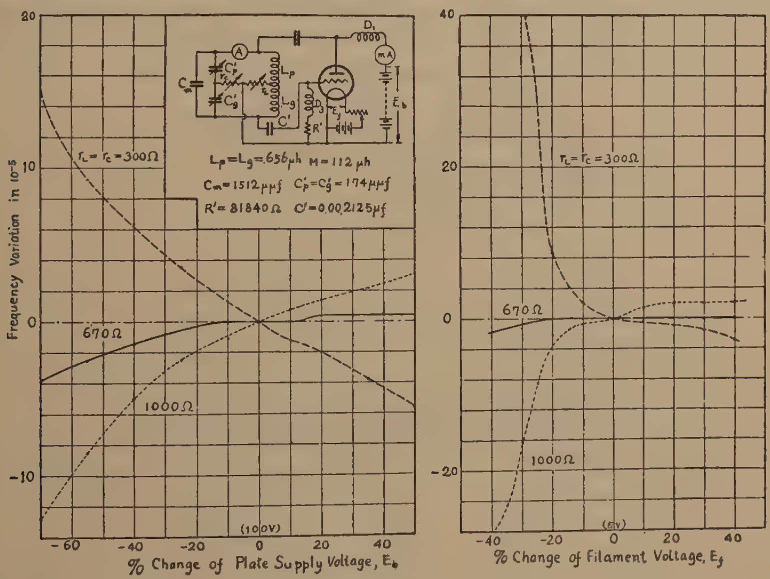

Fig. 9.

of the divided resistances r_L and r_C giving the minimum frequency variation. In some cases mere insertion of this resistance is sufficient for frequency stabilization in a Hartley or a Colpitts circuit. Fig. 8 shows an experimental result in which there are compared the frequency variation characteristics of the three circuits, the Hartley circuit, Colpitts circuit, and the present phase-compensated circuit. It is clearly shown that the phase-compensated oscillator has the variation characteristic which is the mean of the characteristics of Hartley and Colpitts oscillators, and thus produces the stabilizing action. The order of stability obtained in Fig. 8 is within 0.003 per cent for a wide variation in supply voltages.

It is to be noted that the optimum values of the divided resistances for changes in plate voltage are not the same as for changes in filament

voltage. For instance, in the example shown in Fig. 7, optimum values of divided resistances are $r_L=0$ and $r_C=170$ ohms for change of plate supply voltage, while $r_L=597$ ohms and $r_C=1000$ ohms for change of filament voltage, each giving frequency stability of 0.0015 per cent for wide variations in supply voltages. However in practical application, when variation in supply voltages are of the order of ± 10 per cent, we can adopt the mean of these values without causing noticeable instability.

For the sake of convenient adjustment of the circuit as well as for applying the present method to an existing oscillator, a combination of

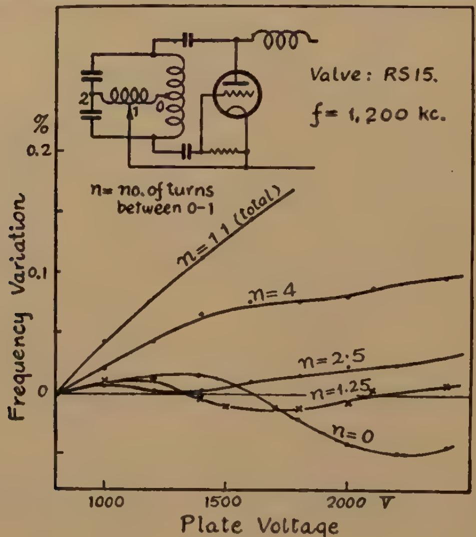

Fig. 10.

main condenser (C_m) and two vernier condensers (C_p', C_g') as shown in Fig. 9 is sometimes to be recommended. Although this circuit is liable to represent the predominant characteristic of a Hartley oscillator, the frequency stability is equally good if adjustment is properly made.

In some cases the bridging resistances may be replaced by an inductance coil having a variable tap, the stabilizing action still being obtained, and this is convenient for application to higher frequencies. An example is shown in Fig. 10, which was obtained with the power valve RS–15 (rated plate voltage 4000v, filament 16v, output 1.5kw).

III. Quartz-Stabilized Oscillator

In an ordinary quartz-oscillator, the power output cannot be increased beyond a certain limit, usually of the order of several watts,

since a higher output gives rise to the overloading of the quartz plate and leads to its damage. The writers attempted to increase the power output of the quartz-controlled oscillator, and developed the method described in the following, the principle of which is illustrated in Fig. 11. An ordinary triode oscillator such as shown in Fig. 11(a) is used, of which the oscillation current I does not vary much under the ordinary working condition with the variation of the capacity C of the oscillator circuit (curve a'). A parallel resonant circuit Z, having a resonance frequency equal to the desired frequency of oscillation, is then introduced into the grid circuit as shown in Fig. 11(b), and when C is varied, the current I diminishes down to zero around the resonance frequency of Z (curve b'), due to the high impedance of Z intercepting the grid ex-

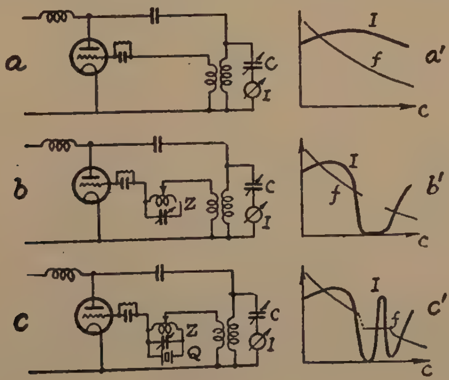

Fig. 11.

citation. After the circuit is adjusted in such a manner, a quartz plate Q accurately cut to give the desired frequency is bridged across the impedance Z as shown in Fig. 11(c), and C is again varied very slowly. Then within the range of C, where the oscillation has once stopped, the current I starts again as shown in curve c' in Fig. 11. In this case the oscillation takes place again owing to the vibration of the quartz which lowers the effective impedance of Z, and the frequency of oscillation is stabilized at the resonance frequency of the quartz. The maximum power output and efficiency obtainable in this stabilized condition are of about the same order as in the ordinary working condition as a self-oscillator.

An experimental result is shown in Fig. 12. A UV-204-A vacuum tube was used at a plate voltage of 1500 volts and a power output of 100 watts. When the vacuum tube was used as an ordinary self-oscillator the oscillation current and the frequency varied with the varia-

tion of the capacity of the oscillation circuit as shown by the broken lines in Fig. 12, but when the grid impedance Z and the quartz plate are utilized and adjusted properly, they varied as shown by the full

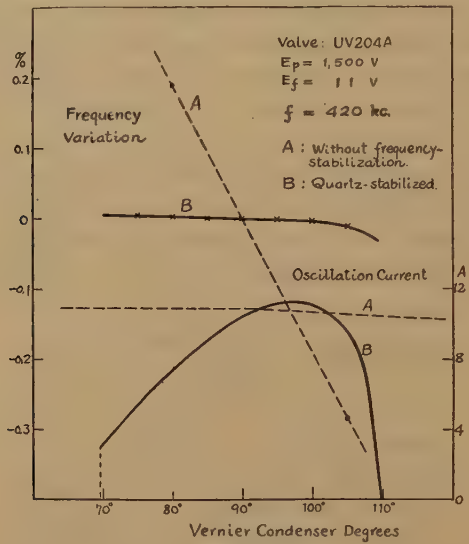

Fig. 12.

lines in the same figure. Thus the frequency of oscillation is stabilized within 0.01 per cent for a variation of the capacity which would give a variation of 0.1 per cent if this stabilizing method were not applied.

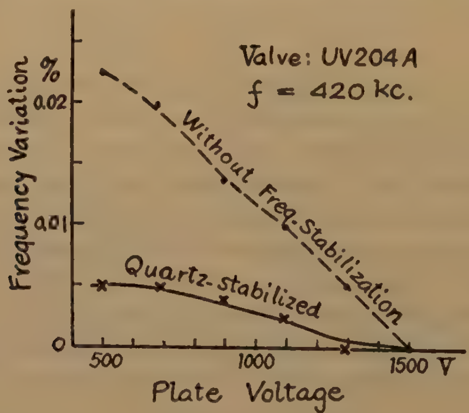

Fig. 13.

Moreover, the oscillation current showed no decrease when compared with the former case. The frequency stability was equally good for the variation of the plate voltage as shown in Fig. 13. The variation of

filament voltage had very little influence on the frequency even in a self-oscillator and no appreciable change was observed in the stabilized state in the whole range of from 7 to 11 volts at which oscillation took place.

By means of the present scheme of quartz stabilization, the power output can be increased to a much higher level than in the ordinary quartz oscillator without causing damage to the quartz plate. This can be attributed to the fact that only a small portion of the grid excitation power, which is already a small fraction of the oscillatory power output, is applied to the quartz plate, while in the ordinary quartz-oscillator a greater part of the output is applied on the vibration of the quartz plate, as the quartz plate forms a part of the main oscillatory circuit through the interelectrode capacity of the valve. Thus an output of the order of 100 watts is easily obtainable, the maximum power output obtained by the writer being of such an order as shown in Table V.

TABLE V

Frequency	Vacuum tube	Plate voltage	Plate input	Power output
745 kc	UV-204-A	2600 v	520 w	280 w
745 kc	MT-7-A	5300 v	1380 w	450 w

In order to take full advantage of the present method, a Curie-cut quartz plate is to be preferred to a 30-degree-cut one, since the latter has been proved to be more liable to damage due to overloading. In using the Curie-cut plate, the upper electrode should be touching the quartz plate in order to obtain reliable operation giving large power output and at the same time to obviate the trouble of sparking in the air gap.

At an oscillating frequency below 1000 kc, no appreciable heating is observed in the quartz plate, when output of the order of 100 watts is being obtained, and the frequency stability can be further assured by placing the quartz plate in a thermostat. But for frequencies higher than this, the heating up of the quartz plate becomes appreciable as the frequency becomes higher, and moreover damage is caused to the quartz plate even at low power output. This is because the oscillating power is fed back to the grid circuit through the grid-to-plate capacity of the triode, and the quartz plate becomes overloaded. In order to overcome this difficulty, a screen-grid tetrode should be utilized, and the writers' experiments proved that the present scheme was successful for giving a power output of 100 watts at frequencies up to 3000 kc. But the heating of the quartz is still appreciable and a special cooling device must be used if the temperature is to be maintained constant. Table VI shows the results of one test.

TABLE VI

Frequency	Vacuum tube	Plate voltage	Plate input	Power output	Temperature rise
2740 kc	UV-861	1400 v 2100 v	170 w 840 w	70 w 250 w	20 degrees C 50 degrees C

Although the stabilized range of frequency as related to a variation of the capacity is narrow compared with the ordinary quartz oscillator, this oscillator can be successfully utilized as a master oscillator, or even as a transmitter directly supplying power to the antenna if the variation of the antenna constants is within the order of 0.1 per cent in frequency. When used in a transmitter in such a manner at a frequency below 1000 kc a frequency stability of 0.01 per cent may be expected over a long period of time if the quartz plate is thermostatically controlled to within 1 degree C.

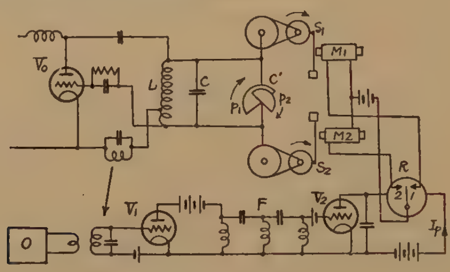

Fig. 14.

IV. Mechanically-Stabilized Oscillator

This method utilizes a mechanical device for stabilizing the frequency of an oscillator which may be of any large power output. A constant-frequency local oscillator of small power is used as the standard of frequency. The circuit arrangement is as shown in Fig. 14. The main oscillator valve V_0 has its accompanying oscillation circuit LC, and the condenser C is shunted by a vernier C' consisting of a pair of movable plates, p_1 and p_2, which are continually subjected to a driving force to rotate them in the directions of the arrows indicated in the figure, although the rotation is, under normal conditions, hindered by the locking pieces S_1 and S_2. A constant-frequency local oscillator of small power O, which should preferably be a quartz oscillator, is used as the frequency standard, and the beat frequency between the fundamental or harmonic frequency of the main oscillator and that of the local oscillator is produced in the detector valve V_1. The beat-frequency current thus produced is passed through the high-pass filter F, and the outgoing voltage is applied to the second detector valve V_2. Then the

rectified direct-current component of the plate current of V_2 shows a sharp increase when the beat frequency passes over the cut-off frequency of the filter, as shown in Fig. 15. The circuit is so adjusted that when the frequency of the oscillator is at the normal value the beat frequency is nearly at the cut-off frequency of the filter and the plate current I_p lies midway between I_1 and I_2 in Fig. 15. A galvanometer relay R is connected in the plate circuit of V_2 and is adjusted such that when I_p is below I_2, contact No. 2 is made, and on the other hand when I_p reaches I_1, contact No. 1 is closed. In either case the circuit comprising the electromagnet M_1 or M_2 is closed and the locking piece S_1 or S_2 releases the wheel and makes p_1 or p_2 rotate in the predetermined direction.

The working principle is very simple: when the oscillator frequency becomes a little higher than normal, relay contact No. 1 is made and the magnet M_1 is energized. The plate p_1 of the vernier condenser then

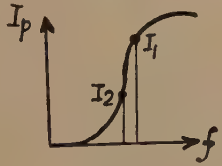

Fig. 15.

begins to rotate in such a direction as to increase the capacity of C' and hence to compensate for the former increase of frequency. When the frequency decreases, relay contact No. 2 is closed and the magnet M_2 is energized and the plate p_2 rotates in a such direction as to decrease the capacity C' so that it counteracts the original decrease in the frequency. As long as the oscillator frequency is in the vicinity of the normal value, the plate current I_p stays between I_1 and I_2, and neither one of these relay contacts is made; hence rotation of the vernier condenser does not take place.

As the frequency variation of a master oscillator is generally slow in character, it can be stabilized most effectively and economically by means of the present method, if the highest precision is not desired. Of course rapid variation in frequency which cannot be compensated completely by the present method may take place, but as it is caused principally by the fluctuation of the supply voltages, it can be eliminated by utilizing the stabilizing scheme already described in the present paper.

The present method has been applied successfully to a transmitter of self-oscillator type of 1-kilowatt power output, operating at a plate

voltage of 10,000 volts. Figs. 16 and 17 show results in which the stabilized condition may be compared with the ordinary case. The variation occurring at a very short period in Fig. 16 was caused by the swinging of the antenna by the wind, and even such a rapid variation

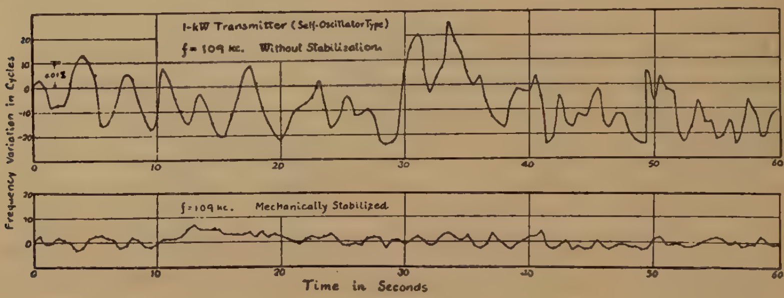

Fig. 16.

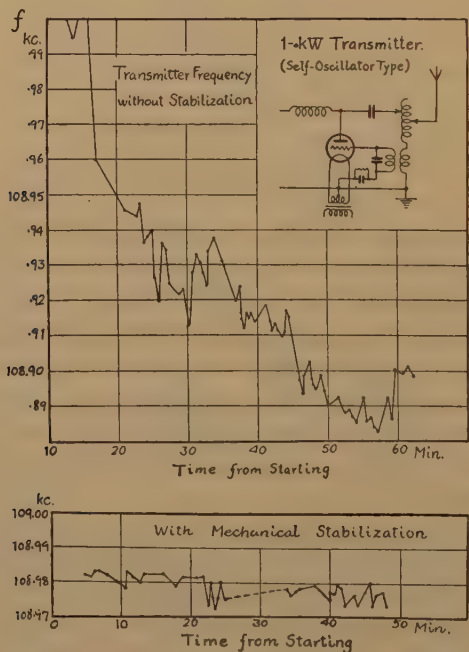

Fig. 17.

is seen to be stabilized to a considerable degree although the speed of rotation of the vernier condenser was not high enough to eliminate it completely. The curve shown in Fig. 17 mainly corresponds to the gradual variation occurring over a long time, amounting to 0.1 per cent

in nearly half an hour, and when the present method was utilized it could be stabilized to within 0.01 per cent.

This scheme of frequency stabilization has its merit in that the oscillator may be of any high power and the auxiliary circuit comprising receiving tubes can be set in a small compartment, so that considerable economy of space as well as of installation cost can be attained. It may be concluded that when this method is applied to a master oscillator which may be followed by only one stage of power amplification, a stability of the order of 0.01 per cent can be obtained, the quartz plate being placed in a thermostat controlled to within 1 degree C.

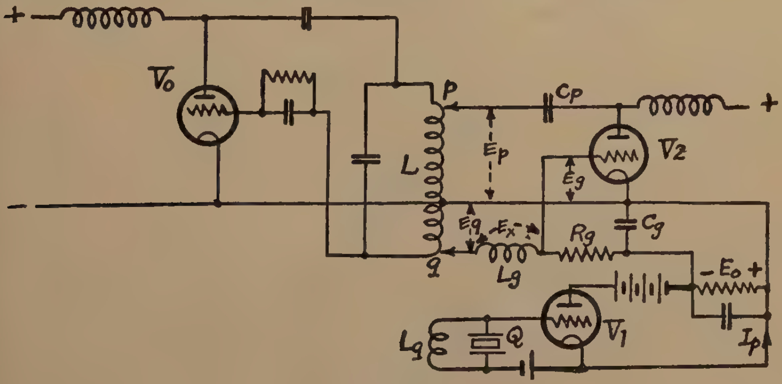

Fig. 18.

V. Vacuum Tube-Stabilized Oscillator

This method makes use of the properties of a vacuum tube whereby it behaves as a pure reactance under proper operating conditions, the reactance value being varied by means of the grid-bias voltage. If such a tube is connected to an oscillator circuit, the oscillating frequency can be controlled with the grid-bias voltage. Hence stabilization can be effected if some means is provided to convert the variation in frequency into that of the grid-bias voltage.

In the present method the circuit arrangement is made as shown in Fig. 18, and a quartz resonator is used as the standard of frequency. The oscillator circuit comprising the oscillator valve V_0 may be of any ordinary type and may deliver any large power output. The stabilizing vacuum tube V_2 which is to act as a reactance, is coupled to the oscillator circuit as shown in the figure, the plate being connected to a point p at one end of the main inductance L through the blocking condenser C_p, and the grid being connected to a point between L_g and Λ_g. The inductance L_g and the resistance Λ_g connected in series are inserted

between the point q on the inductance L and the cathode lead of the vacuum tube through a by-pass condenser C_g. Thus in Fig. 19, which shows the vector relations among various quantities occurring in the vacuum tube V_2, E_p, and E_q are the voltages derived from the points p and q on the inductance L, respectively, and are in opposite phase from each other. E_q is divided into the voltages E_x and E_g applied on

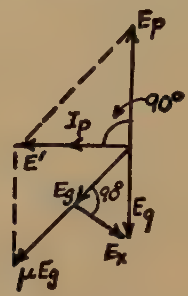

Fig. 19.

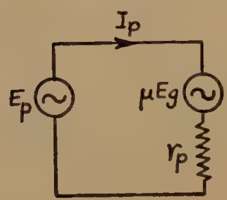

Fig. 20.

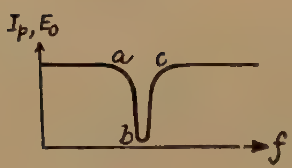

Fig. 21.

L_g and R_g, respectively, and the voltage E_g is applied to the grid of V_2. Under such a condition the plate circuit of the vacuum tube V_2 is equivalent to a circuit shown in Fig. 20, the vacuum tube acting as a source of e.m.f. of μE_g, μ being the amplification constant, and of the internal resistance r_p. Then the plate current of V_2 will be of such an amount as would be flowing under the e.m.f. E' in Fig. 19, which is the vector sum of E_p and μE_g, through the resistance r_p. Hence, $I_p = E'/r_p$, and phase of I_p is the same as that of E'. Then by proper adjustment of the circuit a phase difference of 90 degrees can be obtained between E_p and I_p, and in such a case the vacuum tube is acting

as a pure reactance. By the connection shown in Fig. 18, the vacuum tube is equivalent to a pure capacity of the value:

$$C_e = \frac{I_p}{\omega E_p} = \frac{1}{\omega r_p}\sqrt{\left(\frac{\mu E_g}{E_p}\right)^2 - 1}. \tag{3}$$

Since this equivalent capacity is inversely proportional to r_p, it can be varied by the grid bias voltage, and the increase of the grid voltage in negative sense results in the increase of the resistance r_p and hence decrease of the equivalent capacity, and thus produces an increase in the oscillating frequency in the arrangement shown in Fig. 18.

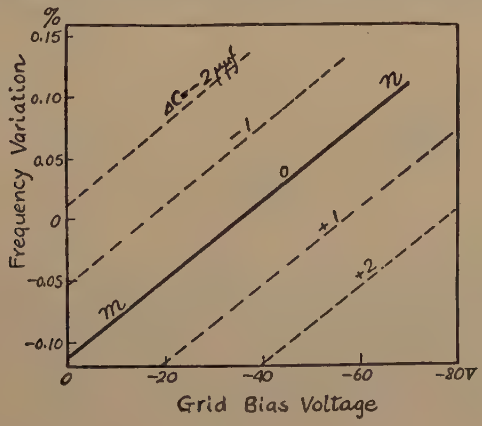

Fig. 22.

In the auxiliary circuit comprising a detector vacuum tube V_1, the coil L_q is loosely coupled to the oscillator, and a quartz plate Q, which is to be used as the frequency standard, is shunted across the inductance. When the oscillating frequency is brought near the resonance frequency of the quartz, the voltage at L_q will be represented by a curve as shown in Fig. 21, and at the resonance point of the quartz the amplitude will show a sudden drop. The steep part of the curve such as the range from a to b is important for the present application. The voltage on L_q is applied to the detector V_1, and the rectified plate current produces a d-c potential difference E_0 at the resistance inserted in the plate circuit. This voltage drop is used as the grid bias voltage on the vacuum tube V_2. Then the variation of the bias voltage will have the same character as the curve abc in Fig. 21, when the oscillator frequency is varied. Adjustment is then made in such a way that when the oscillator is operating at its normal frequency the grid-bias voltage will stay at a point between a and b. Then, if the frequency of the oscillator increases, the negative grid-bias voltage E_0 drops down to-

ward the point b, and the equivalent capacity of the vacuum tube V_2 increases and tends to lower the oscillator frequency and thus counteracts the original increase in the frequency.

Minute analysis of the stabilizing action is to be made by considering the frequency variation characteristic due to the variation of the grid bias E_0 shown in a curve mn in Fig. 22 together with the variation

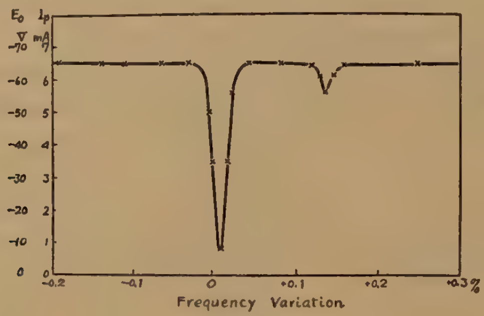

Fig. 23.

characteristic of the grid bias voltage E_0 due to the variation of the oscillating frequency as has been given in Fig. 21, an actual example being shown in Fig. 23. For this purpose the two curves are combined into one diagram as shown in Fig. 24. The points of intersection of these two curves are the equilibrium points, of which the point p_0 in Fig. 24 is a stable point where the operation of the system is depend-

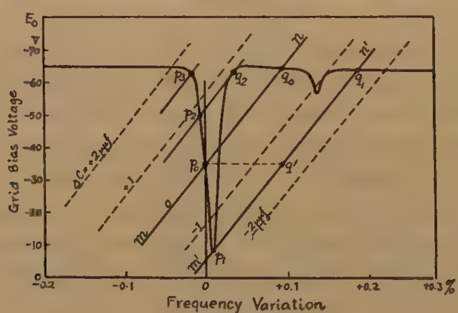

Fig. 24.

able. When the frequency is varied by some cause, for instance by change in the oscillator capacity, shifting of the curve mn takes place to such a position as shown by the group of parallel lines. Thus when mn shifts to a position $m'n'$ the point of intersection shifts accordingly to p_1. If the stabilizing method were not applied, the frequency shift would take place from p_0 to q', as the grid-bias voltage remains unvaried. Thus it is seen that the shift in frequency from p_0 to q' under

the ordinary condition is stabilized to a narrower range of from p_0 to p_1 owing to the present method.

When the change of frequency occurs over a wide range, the steep part of the resonance curve is superseded and the stabilizing action is

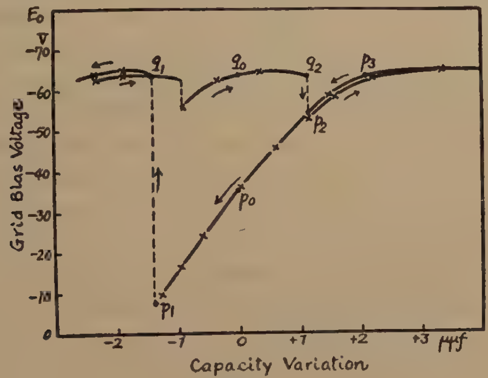

Fig. 25.

lost. Referring to Fig. 24, when the capacity is being increased, the line *mn* shifts toward the left, and the working point moves along $q_1\ q_0\ q_2\ p_2\ p_3$, and from q_2 to p_2 a sudden jump takes place. On the other hand when the capacity is being decreased, the point moves along

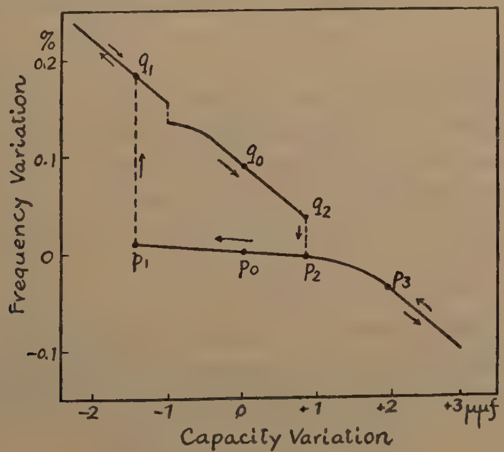

Fig. 26.

$p_3\ p_2\ p_0\ p_1\ q_1$, and a sudden jump occurs from p_1 to q_1. During this shifting of the working point, the oscillating frequency and the grid-bias voltage vary as transcribed in Figs. 25 and 26, the points noted in the figures corresponding to those given in Fig. 24. The range between the points p_1 and p_3 is the most important part at which the present

method of stabilization operates. As long as the capacity variation remains within this limit satisfactory frequency stability of the oscillator may be realized.

The curves shown in the above-mentioned figures have been obtained by using a UV-204-A vacuum tube as an oscillator at a plate voltage of 1500 volts, and another vacuum tube of the same type as a stabilizing vacuum tube. Referring to Fig. 26, it is recognized that frequency is stabilized within ± 0.01 per cent around the point p_0 for the variation in the capacity of $\pm 1\mu\mu$f which corresponds to a variation of 0.1 per cent in this case. The stabilization was equally effective for a variation in the plate voltage as well as the filament voltage. Over a range of from 1200 to 2000 volts of the former and from 8 to 11 volts of the latter, the total variation was within 0.02 per cent.

The power capacity required for the stabilizing vacuum tube V_2 depends on the degree of the frequency variation to be balanced out. In the above example the vacuum tube used was of the same type as the oscillator, and in such a case the same power source can be commonly used for the two vacuum tubes.

The present method is applicable to an oscillator of any large power output, and a frequency variation of rapid nature can effectively be compensated. But strictly speaking, the quartz has a very low decrement and its vibration shows a little time lag, so that very rapid variations occurring with a period of less than say 0.1 second may not be completely stabilized.

The oscillator should be used as a master oscillator in a transmitter and may be followed by only one stage of power amplifier. The degree of stability obtainable depends on the adjustment of the circuits and the capacity of the stabilizing valve, but the order of 0.01 per cent will be easily assured, if the quartz plate is kept constant in temperature within 1 degree C.

Acknowledgment

This work was done under the direction of Mr. E. Yokoyama, Head of the Radio Section of the Laboratory, and, in carrying out the experiments, assistance was rendered by Mr. S. Komatsu and Mr. T. Yanagi.

Appendix

Frequency Formula for Hartley, Colpitts, and "Phase-compensated" Oscillators

(1) Hartley

Applying Kirchhoff's law to the alternating components of voltage and current in Fig. 27, we get

$$i_p r_p = e_p + \mu e_g, \quad i_g r_g = e_g + \frac{e_p}{\nu}$$

$$e_p = e_g - \frac{i_C}{CD} - R_C i_C - R_p i_p + R_g i_g$$

$$e_p = -L_p D(i_p - i_C) - R_{Lp}(i_p - i_C) - R_p i_p + MD(i_g + i_C)$$

$$e_g = -L_g D(i_g + i_C) - R_{Lg}(i_g + i_C) - R_g i_g + MD(i_p - i_C)$$

$$D = \frac{d}{dt}, \quad \frac{1}{D} = \int dt.$$

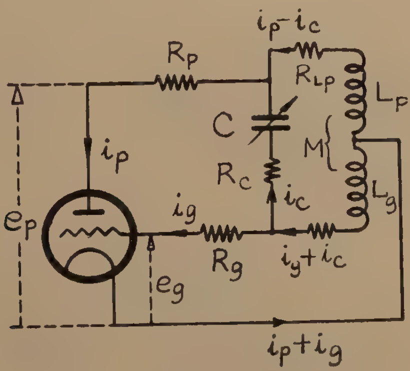

Fig. 27.

Eliminating e_p, e_g, i_g, and i_p, and putting $i_c = I\epsilon^{j\omega t}$, then,

$$-aj\omega^3 - b\omega^2 + cj\omega + d = 0$$

$$\omega^2 = \frac{c}{a} \quad \text{or} \quad = \omega^2 \frac{d}{b}$$

where,

$$a = (L_pL_g - M^2)\left\{\left(1 + \frac{1}{\nu}\right)r_p + (1+\mu)r_g + k(R_p + R_g + R_C)\right\}$$

$$b = (L_p + L_g + 2M)\{r_p(r_g + R_g) + R_p(r_g + kR_g)\} + \frac{k}{C}(L_pL_g - M^2)$$

$$+ R_C\left\{L_p(r_g + kR_g + kR_{Lg}) + L_g(r_p + kR_p + kR_{Lp}) - M\left(\frac{r_p}{\nu} + \mu\nu_g\right)\right\}$$

$$+ (R_{Lp}L_g + R_{Lg}R_p)\left\{\left(1 + \frac{1}{\nu}\right)r_p + (1+\mu)r_g + k(R_p + R_g)\right\}$$

$$c = \frac{L_p}{C}(r_g + kR_g + kR_{Lg}) + \frac{L_g}{C}(r_p + kR_p + kR_{Lp}) - \frac{M}{C}\left(\frac{r_p}{\nu} + \mu r_g\right)$$

$$+\frac{R_C}{k}\left\{(r_p + kR_p + kR_{Lp})(r_g + kR_g + kR_{Lg}) - \frac{\mu}{\nu} r_p r_g\right\}$$

$$+ (R_{Lp} + R_{Lp})\{r_p(r_g + R_g) + R_p(r_g + kR_g)\}$$

$$+ R_{Lp}R_{Lg}\left\{\left(1 + \frac{1}{\nu}\right)r_p + (1 + \mu)r_g + k(R_p + R_g)\right\}$$

$$d = \frac{1}{kC}\left\{(r_p + kR_p + kR_{Lp})(r_g + kR_g + kR_{Lg}) - \frac{\mu}{\nu} r_p r_g\right\}$$

$$k = 1 - \frac{\mu}{\nu}$$

(2) *Colpitts*

In the same manner as (1), we get the frequency formula for the Colpitts oscillator (Fig. 28) as follows:

$$- a'j\omega^3 - b'\omega^2 + c'j\omega + d' = 0$$

$$\omega^2 = \frac{c'}{a'} \quad \text{or} \quad \omega^2 = \frac{d'}{b'}$$

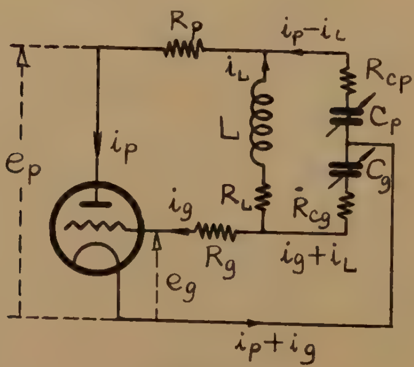

Fig. 28.

where,

$$a' = L\{r_p r_g + r_p(R_g + R_{Cg}) + r_g(R_p + R_{Cp}) + k(R_p + R_{Cp})(R_g + R_{Cg})\}$$

$$b' = L\left\{\left(\frac{r_p}{C_g} + \frac{r_g}{C_p}\right) + \frac{k}{C_p}(R_g + R_{Cg}) + \frac{k}{C_g}(R_p + R_{Cp})\right\}$$

$$+ r_p\{(R_g + R_{Cg})(R_L + R_{Cp}) + R_g R_{Cg}\}$$

$$+ r_g\{(R_p + R_{Cp})(R_L + R_{Cg}) + R_p R_{Cp}\}$$

$$+ r_p r_g(R_L + R_{Cp} + R_{Cg}) + R_{Cp}R_{Cg}\left(\frac{r_p}{\nu} + \mu r_g\right)$$

$$+ k(R_g + R_{Cg})\{R_p(R_{Cp} + R_L) + R_{Cp}R_L\} + kR_g R_{Cg}(R_p + R_{Cp})$$

$$c' = \frac{1}{C_p}\left\{r_p(R_g+R_{Cg})+r_g(R_p+R_{Cp}+R_L)+R_{Cg}\left(\frac{r_p}{\nu}+\mu r_g\right)+r_pr_g\right\}$$

$$+\frac{k}{C_p}\{(R_p+R_L)(R_g+R_{Cg})+R_gR_{Cg}\}$$

$$+\frac{1}{C_g}\left\{r_p(R_g+R_L)+r_g(R_p+R_{Cp})+R_pR_{Cp}\right.$$

$$\left.+R_{Cp}\left(\frac{r_p}{\nu}+\mu r_g\right)+r_pr_g\right\}$$

$$+\frac{k}{C_g}\{(R_p+R_{Cp})(R_g+R_L)+R_pR_{Cp}\}+\frac{kL}{C_pC_g}$$

$$d' = \frac{1}{C_pC_g}\left\{k(R_p+R_g+R_L)+r_p\left(1+\frac{1}{\nu}\right)+r_g(1+\mu)\right\}$$

$$= 1-\frac{\mu}{\nu}.$$

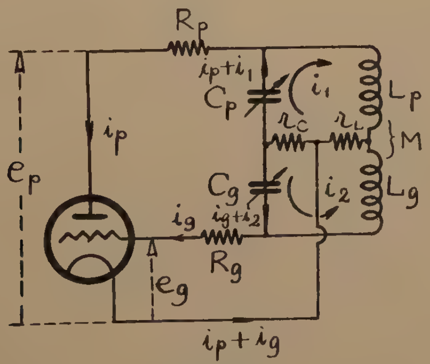

Fig. 29.

(3) *"Phase-Compensated" Oscillator*

Applying Kirchhoff's law, we get six fundamental equations for the circuit of Fig. 29 as follows:

$$i_pr_p = e_p + \mu e_g$$

$$i_gr_g = e_g + \frac{e_p}{\nu}$$

$$e_p = -r_C(i_p+i_g+i_1+i_2) - \frac{i_p+i_1}{C_pD} - R_pi_p$$

$$e_g = -r_C(i_p + i_g + i_1 + i_2) - \frac{i_g + i_2}{C_g D} - R_g i_g$$

$$-r_C(i_p + i_g + i_1 + i_2) - \frac{i_p + i_1}{C_p D} - L_p D i_1 - r_L(i_1 + i_2) + M D i_2 = 0$$

$$-r_C(i_p + i_g + i_1 + i_2) - \frac{i_g + i_2}{C_g D} - L_g D i_2 - r_L(i_1 + i_2) + M D i_1 = 0$$

Eliminating e_p, e_g, i_p, i_g, and i_1, and putting $i_2 = I\epsilon^{j\omega t}$, we get the following biquadratic equation for the frequency of a "phase-compensated" oscillator.

$$p\omega^4 - qj\omega^3 - s\omega^2 + uj\omega + v = 0$$

$$\omega^2 = \frac{s \pm \sqrt{s^2 - 4pv}}{2p}$$

or

$$\omega^2 = \frac{u}{q}$$

where,

$$p = (L_p L_g - M^2)\left\{\frac{1}{k}(r_p + kR_p)(r_g + kR_g) + r_C(r_p + r_g + kR_p + kR_g)\right.$$
$$\left. + r_C\left(\frac{r_p}{\nu} + \mu r_g\right) - \frac{1}{k}\cdot\frac{\mu}{\nu} r_p r_g\right\}$$

$$q = (L_p L_g - M^2)\left\{kr_C\left(\frac{1}{C_p} + \frac{1}{C_g}\right) + \frac{r_g + kR_g}{C_p} + \frac{r_p + kR_p}{C_g}\right\}$$
$$+ (L_p + L_g + 2M)\left\{\frac{1}{k}(r_p + kR_p)(r_g + kR_g)(r_C + r_L)\right.$$
$$\left. + r_C r_L(r_p + r_g + kR_p + kR_g) + r_C r_L\left(\frac{r_p}{\nu} + \mu r_g\right)\right\}$$

$$s = \frac{k}{C_p C_g}(L_p L_g - M^2)$$
$$+ r_C\left(\frac{1}{C_p} + \frac{1}{C_g}\right)\left\{(r_p + kR_p)L_g + (r_g + kR_g)L_p - \left(\frac{r_p}{\nu} + \mu r_g\right)M\right\}$$
$$+ r_L(L_p + L_g + 2M)\left\{kr_C\left(\frac{1}{C_p} + \frac{1}{C_g}\right) + \left(\frac{r_p + kR_p}{C_g} + \frac{r_g + kE_g}{C_p}\right)\right\}$$
$$- \frac{\mu}{k\nu} r_p r_g\left(\frac{L_p}{C_g} + \frac{L_g}{C_p}\right)$$

$$u = \left(\frac{1}{C_p} + \frac{1}{C_g}\right)\left\{\frac{1}{k}(r_p + kR_p)(r_g + kR_g)(r_C + r_L)\right.$$

$$+ r_C r_L(r_p + r_g + kR_p + kR_g) + r_C r_L\left(\frac{r_p}{\nu} + \mu r_g\right)$$

$$- \frac{\mu}{k\nu} r_p r_g (r_C + r_L)$$

$$+ \frac{1}{C_p C_g}\left\{(r_p + kR_p)L_g + (r_g + kR_g)L_p - \left(\frac{r_p}{\nu} + \mu r_g\right)M\right\}$$

$$+ \frac{k r_L}{C_p C_g}(L_p + L_g + 2M)$$

$$v = \frac{1}{C_p C_g}\left\{\frac{1}{k}(r_p + kR_p)(r_g + kR_g) + r_L(r_p + r_g + kR_p + kR_g)\right.$$

$$\left. + r_L\left(\frac{r_a}{\nu} + \mu r_g\right) - \frac{\mu}{\nu k} r_p r_g\right\}$$

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ELIMINATION OF HARMONICS IN VACUUM TUBE TRANSMITTERS*

By

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Summary—*A brief description is first given on the present state of technique concerning the suppression of harmonics radiated from vacuum tube transmitters. A simple method of eliminating the strongest one of the harmonics is then suggested, in which a parallel resonant circuit having its resonant frequency slightly lower than that of the harmonic to be eliminated is connected to the plate circuit and coupled to the main oscillatory circuit. The method has been found effective experimentally as well as theoretically.*

THE reduction of harmonics produced in radio transmitters has long been one of the important problems in radio communication, as the harmonic radiation causes interference to other stations. In the present paper the writer outlines briefly the methods previously known of reducing harmonic radiation and suggests a simple method of eliminating the strongest one of the harmonics. Vacuum tube transmitters solely are considered in this paper.

The production of harmonics in a vacuum tube transmitter is caused by the distored plate current which is generally composed of a series of intermittent peak waves occurring at the frequency of the transmitted wave, and this is inevitable for the efficient operation of a tube. The harmonic components of the plate current thus produced are carried to the antenna through the oscillator circuit, but the harmonic content in the antenna current is remarkably smaller than that in the plate current on account of the selectivity of the resonant circuits. The harmonic content can be further reduced by making a proper choice of the coupling system between the intermediate and the antenna circuits.

These facts have already been revealed by some investigators[1] and various methods of suppressing the harmonics have been adopted in the existing transmitters. The present writer draws the following conclusions from the previous investigations and the present practice in the design of transmitters:

(1) The operation of a vacuum tube in a state of impulse excitation

* Decimal classification: R146. Manuscript originally received by the Institute, June 24, 1931. Distributed at U.R.S.I., Copenhagen, 1931. Printed by National Research Council of Japan.

[1] R. V. Hansford and H. Falkner, "Some notes on design details of a high-power radio-telegraphic transmitter using thermionic valves," *J.I.E.E., Wireless Proc.*, **2**, 10, March, 1927.

should not be carried to an extreme, for this increases the harmonic radiation while contributing little toward raising the working efficiency.

(2) An intermediate resonant circuit should preferably be used between the plate and the antenna circuits. This can be further effected by the use of more than two resonant circuits.

(3) The decrement of the antenna and the intermediate circuits should be as small as technically admissible.

(4) In a parallel resonant circuit inserted in the plate circuit, the harmonic currents flow mainly in the capacitive leg and very little in the inductive leg, so that the power to the antenna should preferably be drawn from the latter.

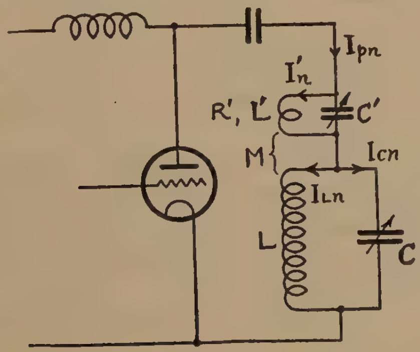

Fig. 1

(5) For the coupling of two resonant circuits capacitive coupling is much more effective in suppressing harmonics than inductive coupling.

(6) A wave filter of low-pass type may advantageously be used in the feeder line leading to the antenna. Series or parallel resonant circuits tuned to the frequency of a harmonic may be utilized in a transmitter circuit in such a way as to prevent the harmonic current from flowing into the antenna circuit.[2]

(7) Elimination of even harmonics can be effected by the push-pull connection of two tubes in the last stage of a transmitter.

Generally speaking, the harmonic currents of higher orders are relatively weak in intensity and they may easily be reduced by some simple means, but those of lower orders, especially the second and the third harmonic, are strong in intensity and much difficulty is experienced in bringing them down to a very low level as their frequencies are relatively close to the fundamental. It may, therefore, be recognized that

[2] F. Banneitz, "Taschenbuch der drahtlosen Telegraphie und Telephonie," p. 725, 1927.

great advantage will be obtained if a simple method is known for the elimination of the strongest ones of the harmonics.

With this object in view the writer has devised the following method, the principle of which lies in the fact that the harmonics in the plate circuit are much stronger than those in the inductive leg of a resonant circuit inserted, and the latter can, therefore, be balanced out by the former, utilizing a suitable coupling between the two circuits.

The circuit arrangement is as shown in Fig. 1. In the plate circuit of the vacuum tube, which may be either a self-oscillator or a power amplifier, a parallel resonant circuit $L'C'$ is inserted and the inductance L' is coupled to the main resonant circuit LC. The resonant frequency of $L'C'$ should be a little lower than the frequency of the harmonic to be eliminated. A balanced condition may then be found by the adjustment of M or C', which is to bring the harmonic current in the inductive leg down to a minimum.

The necessary condition can be deduced as follows. Let the fundamental frequency be ω and assume that the n-th harmonic current in the inductance L could have been eliminated. Then referring to Fig. 1 which shows the distribution of the n-th harmonic currents,

$$I_{Ln} = 0$$

hence,

$$I_{Cn} = I_{pn}$$

then,

$$\begin{cases} \dfrac{I_{pn}}{jn\omega C} = jn\omega M I_n' \\[2ex] \dfrac{I_{pn} - I_n'}{jn\omega C'} = (R' + jn\omega L')I_n \end{cases}$$

$$\therefore \quad n^2\omega^2 CM = n^2\omega^2 C'L' - jn\omega C'R' - 1$$

putting

$$\omega^2 = \frac{1}{LC}$$

$$n^2\frac{M}{L} = \left(n^2\frac{C'L'}{CL} - 1\right) - jn\omega C'R'.$$

Assuming that the condition

$$R' \ll \frac{n}{\omega C'}\left(\frac{C'L'}{CL} - \frac{1}{n^2}\right)$$

is fulfilled,

$$\frac{M}{L} = \frac{C'L'}{CL} - \frac{1}{n^2}. \tag{1}$$

This gives the condition at which the n-th harmonic can be eliminated in the inductive leg of the resonant circuit.

In order that the above assumption holds, the circuit $C'L'$ must be detuned below the resonant condition at the n-th harmonic, because at the resonant condition the term in the parentheses becomes zero. The expression (1) gives the necessary coupling for a certain value of $C'L'$ which can be chosen at a convenient value.

Table I gives the experimental results obtained in a self-oscillator, in which harmonic currents are expressed in terms of relative amplitudes.

TABLE I

Valve UV—204 A		Second harmonic current	Working condition		
			Plate voltage	Plate current	Oscillation current
Fundamental frequency 1042 kc	Without $L'C'$	10	2000 V	0.09 A	13.5 A
	With $L'C'$ tuned to 2nd harmonic; $M=0$	3	2000	0.08	13.5
	With writer's method	1	2000	0.05	11.8
Fundamental frequency 375 kc	Without $L'C'$	100	1500	0.08	15
	With writer's method	4			

In case of a self-oscillator, grid current is also responsible for the production of harmonics, but this effect may also be cancelled by the present method.

An actual test was made by applying the method to a transmitter of the Tokyo Central Broadcasting Station, and the result shown in Table II was obtained. The insertion of a balancing circuit in the plate circuit of the last stage amplifier had no effect on the working condition and power output of the transmitter.

TABLE II

Fundamental frequency	Extent of reduction of the second harmonic current		Working condition			
					Oscillation current	
	Intermediate circuit	Antenna	Plate voltage	Plate current	Int. Cct.	Antenna
590 kc	3/100	1/100	9600 V	2.1 A	14.5 A	20 A

Another balancing condition may be obtained when C' is not used; putting $C'=0$ in (1), the required condition is obtained as follows:

$$\frac{M}{L} = -\frac{1}{n^2}. \tag{2}$$

In this case M is negative and the sense of coupling must be re-

versed. An experimental result is shown in Fig. 3, which was obtained by a circuit as shown in Fig. 2. The adjustment of M was made by varying the number of turns of L', and at a proper value of L' the second harmonic is eliminated as shown in Fig. 3.

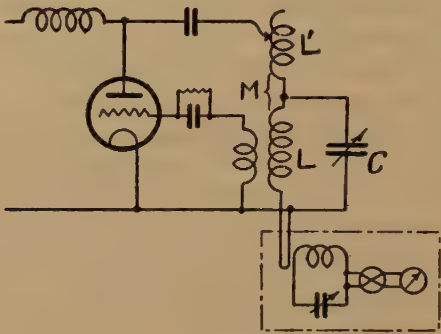

Fig. 2

In the former case, if $C'L'/CL$ is large compared with $1/n^2$, the harmonics other than that balanced out by the method can also be reduced to some extent, but in the latter case only the one of the harmonics is reduced and the others are not.

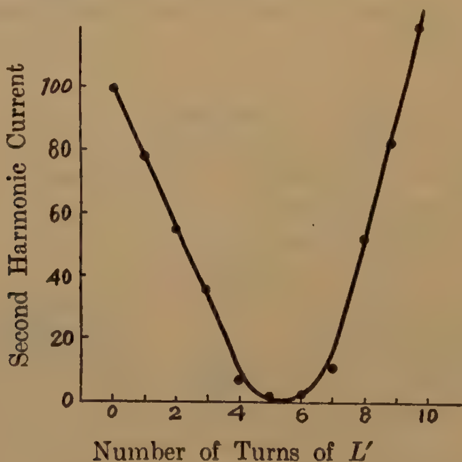

Fig. 3

In both cases the operating condition of the tube is sometimes affected due to the insertion of extraneous impedance in the plate circuit, but it only alters the plate circuit impedance and the power absorbed in this circuit is not appreciable, and hence the normal output can be obtainable by adjusting the main circuit impedance.

The writer is of the opinion that the harmonics of a transmitter can be reduced to a practically admissible level if the present method is adopted for the elimination of the second harmonic and an intermediate-circuit which is capacitively coupled to the antenna circuit is utilized for further reduction of the harmonics, especially of higher orders.

Acknowledgment

The present work was done under the direction of Mr. Yokoyama, head of the Radio Section, and in making the experiment at the Tokyo Central Broadcasting Station the writer was much indebted to the members of the Station.

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CHARACTERISTICS OF AIRPLANE ANTENNAS FOR RADIO RANGE BEACON RECEPTION*

By
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Summary—*This paper gives the results of an investigation on the characteristics of airplane receiving antennas to determine whether an antenna arrangement could be devised which would have all the desirable electrical properties of the vertical pole antenna and yet be free from the mechanical difficulties encountered in the use of the pole antenna. The antennas studied include the inclined antenna with both forward and backward inclination, the horizontal dipole antenna, the horizontal L antenna, the horizontal V antenna, the inclined V antenna, the symmetrical transverse T antenna, and the symmetrical longitudinal T antenna. A theoretical treatment is given which enables the voltage induced by a radio range beacon transmitting station to be calculated for any receiving antenna in space. This theoretical analysis is used to determine the received voltage, course error, and localizing effect for each of the antenna types studied. An experimental study was also made to check the theoretical analysis. The results obtained by experiment check very well with the theoretical predictions for each type of antenna. The symmetrical transverse T antenna and the symmetrical longitudinal T antenna, with vertical lead-in portions, are both found to fulfill the desired requirements. Neither of these antennas show any course errors, and give the same received voltage as the vertical pole antenna having much greater actual height, thus reducing the mechanical troubles caused by vibration and ice formation.*

I. Introduction

SINCE the advent of the radio range beacon, the vertical pole antenna has been widely used because of its nondirectional properties and consequent freedom from course errors in radio range beacon reception. The vertical pole antenna, however, is subject to considerable mechanical trouble because of vibration and ice formation, and for this reason, an antenna free from these mechanical difficulties and yet having all the desirable electrical properties of the simple pole antenna would be preferable.

Accordingly, a study of various types of airplane antennas was begun, a theoretical analysis being checked by experimental observations in the air and on the ground. A number of antenna arrangements were included in this study. For each type of antenna, the tests in the air included observation of the received voltage, the localizing effect or variation of the received voltage in the immediate vicinity of the

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beacon tower, and the course error as observed by circling the beacon. These were compared directly with the results obtained using the vertical pole antenna. The types studied included the inclined antenna, with both forward and backward inclination (one example of the latter being the trailing wire antenna), the horizontal dipole antenna, the horizontal L antenna, the horizontal V antenna, the inclined V antenna, the symmetrical transverse T antenna, and the symmetrical longitudinal T antenna. Fig. 1 shows experimental installations of these antennas on an airplane. The results of this experimental work check the theoretical analysis with sufficient accuracy for the desired purpose, although exact quantitative data are almost impossible to obtain in the air because of the difficulties incurred in any attempt to measure the geometrical quantities involved.

II. Theoretical Analysis

It is found convenient to treat the subject by the use of vector analysis. A somewhat similar treatment, using a trigonometric method, has been given by William H. Murphy of the Signal Corps.[1] In this work, as in Murphy's, relative rather than absolute field intensities and received voltages are desired. Consequently, all constants are considered equal to unity and omitted from the vector equations. The airplane is also considered to be at unit distance from the transmitter.

The transmitting loop antenna is taken in the $\bar{j}$-$\bar{k}$ plane, with its center at O (Fig. 2), so that the $\bar{i}$-axis coincides with the axis of the loop antenna. The airplane is located at the point Q whose position vector is $\bar{r}$, and is flying in the direction indicated by the line marked "line of flight." The line of flight is assumed to lie in a plane parallel to the $\bar{i}$-$\bar{j}$ plane. The vector $\bar{a}$ is a unit vector in the direction of the antenna, pointing away from Q. α_1 is the angle between the axis of the loop antenna ($\bar{i}$-axis) and the projection of the line of sight (or the vector $\bar{r}$) on the $\bar{i}$-$\bar{j}$ plane. α_2 is the angle between the line of flight and the projection of the line of sight on the horizontal plane containing the line of flight. β_1 is the angle of elevation of the airplane. β_2 is the angle of inclination of the antenna with respect to the horizontal plane. These angles are taken as positive when measured in the directions indicated in the figure. The antenna on the airplane is here considered to be in the vertical plane containing the longitudinal axis of the fuselage. This limitation may be removed, however, and the antenna considered in any position, by replacing α_2 in the following equations by $\alpha_2+\alpha_2'$,

[1] W. H. Murphy, "Space characteristics of antennae," *Jour. Frank. Inst.*, 201, 411–429; April, 1926; 203, 289–312; February, 1927.

where α_2' is the angle between the longitudinal axis of the airplane and the horizontal projection of the antenna.

The effective value of the electric field at Q, unit distance from the transmitter, may be written, using Gibb's notation[2]

$$\bar{E}_1 = \bar{r} \times \bar{\imath}. \tag{1}$$[2]

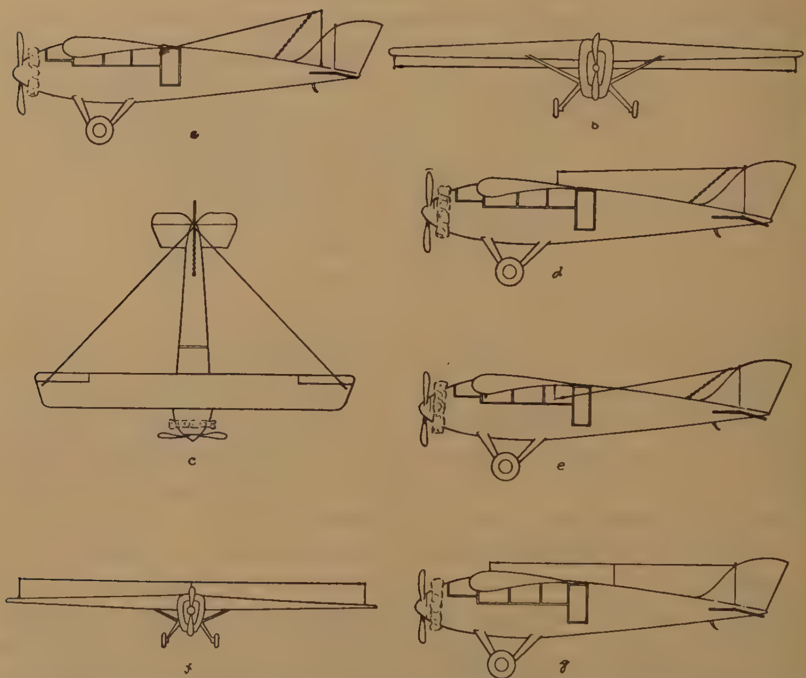

Fig. 1—Experimental installations of antennas studied. *a*—inclined antenna; *b*—horizontal dipole antenna (half of this was used for a horizontal *L* antenna); *c*—top view, *V* antenna; *d*—side view, horizontal *V* antenna; *e*—side view, inclined *V* antenna; *f*—symmetrical transverse *T* antenna; *g*—symmetrical longitudinal *T* antenna.

The effective voltage induced in the receiving antenna by this field vector is

$$E_1 = \bar{E}_1 \cdot \bar{a} = \bar{r} \times \bar{\imath} \cdot \bar{a} = [\bar{r}\bar{\imath}\bar{a}]. \tag{2}$$

If the vectors $\bar{r}$ and $\bar{a}$ are evaluated in terms of their scalar components r_1, r_2, r_3, a_1, a_2, and a_3, the value of this expression may be computed. However, (2) is readily simplified still further:

[2] For a proof of this equation see H. Diamond and G. L. Davies, "Characteristics of airplane antennas for radio range-beacon reception," appendix of Research Paper 313, *Bureau of Standards Journal of Research*, 6, 901–916; May, 1913.

$$E_1 = [\bar{r}\bar{\imath}\bar{a}] = \begin{vmatrix} r_1 & r_2 & r_3 \\ 1 & 0 & 0 \\ a_1 & a_2 & a_3 \end{vmatrix} = -\begin{vmatrix} r_2 & r_3 \\ a_2 & a_3 \end{vmatrix} = a_2 r_3 - r_2 a_3. \tag{3}$$

Similarly, for a second loop antenna in the $\bar{\imath}$-$\bar{k}$ plane, with its axis coinciding with the $\bar{\jmath}$-axis,

$$\bar{E}_2 = \bar{\jmath} \times \bar{r} \tag{4}$$

and,

$$E_2 = \bar{\jmath} \times \bar{r} \cdot \bar{a} = [\bar{\jmath}\bar{r}\bar{a}] \tag{5}$$

$$E_2 = \begin{vmatrix} 0 & 1 & 0 \\ r_1 & r_2 & r_3 \\ a_1 & a_2 & a_3 \end{vmatrix} = \begin{vmatrix} r_1 & r_3 \\ a_1 & a_3 \end{vmatrix} = a_1 r_3 - r_1 a_3. \tag{6}$$

Now the components r_1, r_2, r_3, a_1, a_2, and a_3 of the unit vectors $\bar{r}$ and $\bar{a}$ may be expressed in any system of coördinates, but the most convenient form is that of the spherical coördinates shown in Fig. 2. In this system, and since the vectors are unit vectors (i.e., $|\bar{r}| = |\bar{a}| = 1$),

$$\left.\begin{aligned} r_1 &= \cos\alpha_1 \cos\beta_1 \\ r_2 &= \sin\alpha_1 \cos\beta_1 \\ r_3 &= \sin\beta_1 \end{aligned}\right\} \tag{7}$$

and,

$$\left.\begin{aligned} a_1 &= \cos(\alpha_1 + \alpha_2)\cos\beta_2 \\ a_2 &= \sin(\alpha_1 + \alpha_2)\cos\beta_2 \\ a_3 &= \sin\beta_2 \end{aligned}\right\} \tag{8}$$

On substitution of these values, (3) and (6) become

$$E_1 = a_2 r_3 - r_2 a_3 = \sin(\alpha_1 + \alpha_2)\sin\beta_1\cos\beta_2 - \sin\alpha_1\cos\beta_1\sin\beta_2 \tag{9}$$

$$E_2 = a_1 r_3 - r_1 a_3 = \cos(\alpha_1 + \alpha_2)\sin\beta_1\cos\beta_2 - \cos\alpha_1\cos\beta_1\sin\beta_2 \tag{10}$$

These two equations form the basis of the remainder of the theoretical analysis.

A course is obtained whenever $E_1 = E_2$. Thus, for the general antenna shown in Fig. 2, a course is obtained when

$$\begin{aligned} &\sin(\alpha_1 + \alpha_2)\sin\beta_1\cos\beta_2 - \sin\alpha_1\cos\beta_1\sin\beta_2 \\ &\qquad = \cos(\alpha_1 + \alpha_2)\sin\beta_1\cos\beta_2 - \cos\alpha_1\cos\beta_1\sin\beta_2 \end{aligned} \tag{11}$$

or when,

$$\tan\alpha_1 = \frac{\tan\beta_1(\cos\alpha_2 - \sin\alpha_2) - \tan\beta_2}{\tan\beta_1(\sin\alpha_2 + \cos\alpha_2) - \tan\beta_2} \tag{12}$$

where α_1 is the course angle measured from axis of loop antenna 1.

If the currents in the two loop antennas are equal, the true courses bisect the angles between the transmitting loop antennas. Assuming equal currents, then, the course error for any type of receiving antenna may be found. If the course error is here defined[3] by e, where

$$e = \alpha_1 - \pi/4 \tag{13}$$

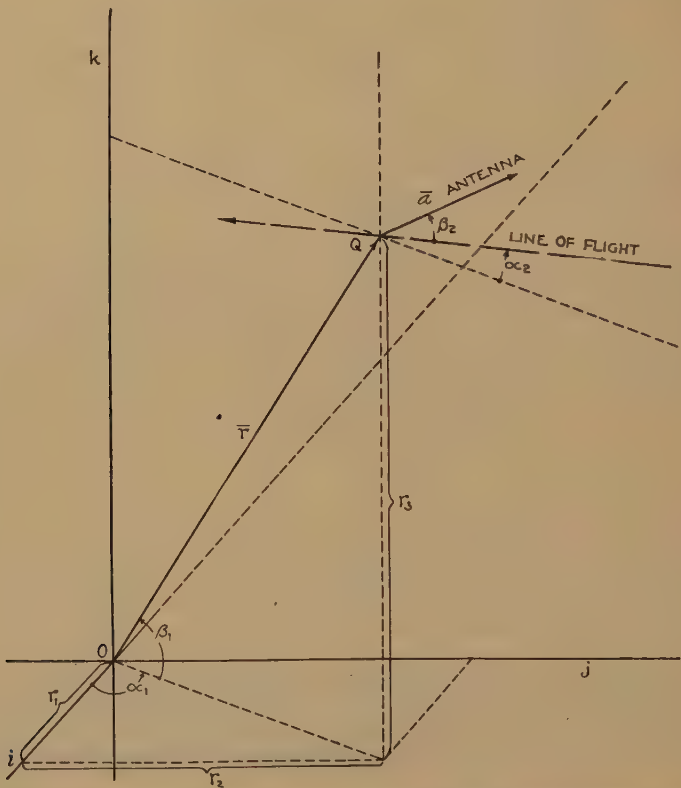

Fig. 2—Diagram showing quantities used in the theoretical analysis.

then,

$$\tan e = \tan(\alpha_1 - \pi/4) = \frac{\tan \alpha_1^{-1}}{1 + \tan \alpha_1}. \tag{14}$$

Combining (12) and (14)

$$\tan e = \frac{\sin \alpha_2 \tan \beta_1}{\tan \beta_2 - \cos \alpha_2 \tan \beta_1}. \tag{15}$$

[3] This is the usual way of defining course error in connection with flight on a radio range beacon.

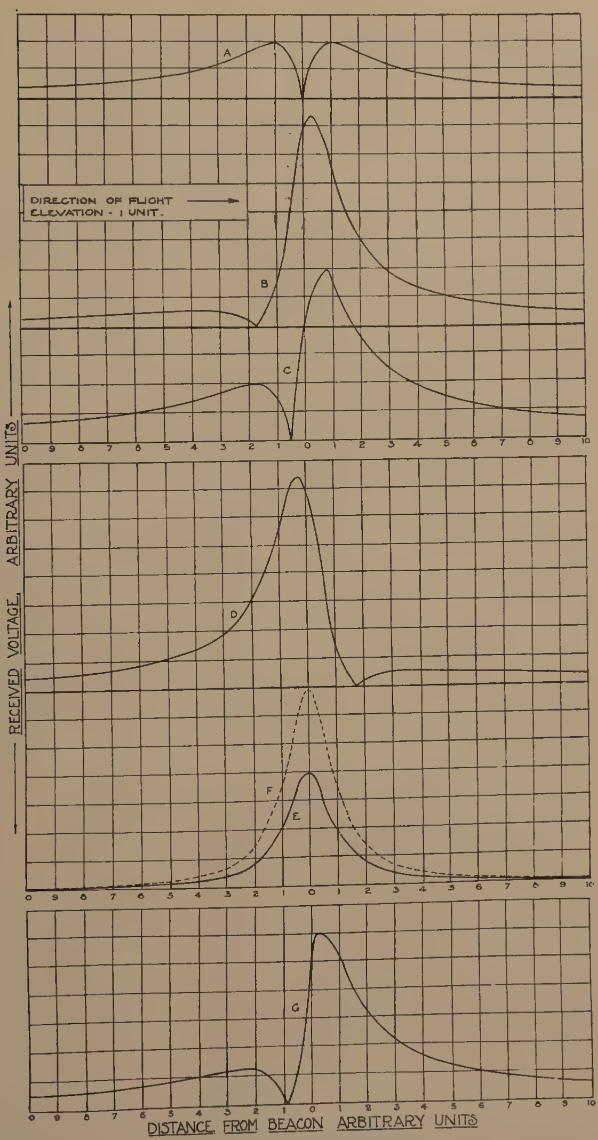

Fig. 3—Graphs of localizing effects for antennas studied. *A*—vertical antenna or symmetrical *T* antenna of either longitudinal or transverse type; *B*—inclined antenna ($\beta_2 = 30°$); *C*—inclined antenna ($\beta_2 = 60°$); *D*—trailing wire antenna ($\beta_2 = -30°$); *E*—horizontal *L* antenna; *F*—horizontal *V* antenna (angle between each wire and fuselage center line $= 30°$); *G*—inclined *V* antenna, angle between each wire and fuselage center line $= 30°$, angle of inclination $= 45°$.

Mathematically, it is possible to make this course error zero in several ways: $\beta_1 = 0$, $\alpha_2 = 0$, or $\beta_2 = \pi/2$. The last is the only practicable method—heretofore accomplished through the use of a vertical antenna. However, a consideration of the physical significance of (11) shows that the course error is introduced by the horizontal component of the field vector, for (11) can be true for $\alpha_1 = 45$ deg. (i.e., $e = 0$) only if $\alpha_2 = 0$. Since α_2 can be considered as a generalized direction angle of the vector sum of the horizontal projections of the antenna, it follows that α_2 can be zero for all directions of flight only if the vector sum of the horizontal projections of the antenna is zero. Obviously, the horizontal projection of the antenna is affected only by the horizontal component of the field vector, so that it is evident that the course error is introduced by this horizontal component. The horizontal component of the field vector cannot be eliminated (except by changes in the transmitting antenna, which are not convenient), so that it becomes necessary to construct the receiving antenna so that it will not respond to this horizontal component: that is, as noted above, the vector sum of the horizontal projections of the receiving antenna must be zero. Since the use of flat-top portions is thus permissible so long as they fulfill these conditions, it becomes possible to design antennas having smaller actual height than that of the usual pole antenna, and yet having the same effective height. Such an antenna would give the same received voltage as the vertical pole antenna, and would likewise be immune to course errors.

III. Application of Theory to Specific Antenna Structures

The theory is readily applied to indicate the results obtainable from several types of simple antennas. In each case, it is desirable to determine the localizing effect (the behavior of the signal in the immediate vicinity of the beacon), the course error, and the relative received voltage. The accompanying graphs (see Fig. 3) show the localizing effect for the antennas considered.

A. Vertical Antenna

$$\beta_2 = \frac{\pi}{2}. \tag{16}$$

1. Localizing effect: Substituting this condition in (9) and (10) and adding, the total received voltage is

$$E = E_1 + E_2 = -\sin\alpha_1 \cos\beta_1 - \cos\alpha_1 \cos\beta_1. \tag{17}$$

This equation is strictly correct only when the two loop antennas are energized simultaneously and in the same time phase (for example, in the case of the two-course visual type beacon). However, the results secured in the following analysis are independent of this limitation, although the equations used in securing these results are somewhat different if the loop antenna currents are displaced from each other in time phase, or if the loop antennas are energized separately as in the case of the aural-type beacon.

The value of E in (17) becomes zero when

$$\beta_1 = \frac{\pi}{2}. \tag{18}$$

Thus, this antenna gives a zero signal zone directly over the beacon (Fig. 3A).

2. Course error: Substituting $\beta_2 = \pi/2$ in (15),

$$e = 0. \tag{19}$$

3. Received voltage: In (17), α_1 may have any value, but is constant for any particular line of flight over the beacon, so that E is proportional to $\cos \beta_1$. Thus, at great distances, where β_1 is small, the vertical antenna gives a relatively large received voltage.

B. Inclined Antenna

The antenna is considered in the same vertical plane as the fuselage, either inclined forward ($\beta_2 > 0$), or trailing wire ($\beta_2 < 0$). The general equations given above apply to this case.

1. Localizing effect: This antenna gives both a zero signal zone and a maximum signal zone, one occurring before the beacon is reached and the other after it has been passed (see (9) and (10)). If β_2 is positive, the zero signal zone occurs first (Figs. 3B and 3C), while if β_2 is negative, the maximum occurs first (Fig. 3D). The magnitude of β_2 determines the distances of these two zones from the beacon. The sharpness of the zero and the magnitude of the maximum are also dependent on the value of β_2. As β_2 is decreased, the zero signal zone recedes from the beacon, and the maximum increases in value and approaches the beacon. In the limiting case ($\beta_2 = 0$), the maximum occurs directly over the beacon and the zero signal zone disappears (recedes to infinity).

2. Course error: This is given by (15). Thus, if the airplane is flying directly toward or away from the beacon ($\alpha_2 = 0$)

$$e = 0. \tag{20}$$

If the airplane is circling the beacon (i.e., $\alpha_2 = \pi/2$),

$$\tan e = \frac{\tan \beta_1}{\tan \beta_2}. \tag{21}$$

If $\alpha_2 = \pi/4$ and $\beta_2 = \pi/4$

$$\tan e = \frac{\sqrt{2} \tan \beta_1}{2 - \sqrt{2} \tan \beta_1}. \tag{22}$$

In (15), when $\tan \beta_2 - \cos \alpha_2 \tan \beta_1$ becomes very small, $\tan e$ becomes very large and e becomes nearly equal to $\pi/2$.

3. Received voltage: If β_2 is large, (greater than $\pi/3$), the received voltage in an inclined antenna is very nearly the same as that in a vertical antenna of the same length. However, if β_2 is small, the inclined antenna is markedly inferior to the vertical antenna at great distances from the beacon, although there is a region, in the vicinity of the maximum signal zone mentioned above, where the inclined antenna is superior to the vertical antenna. This is a small region close to the beacon, and of little value practically, except for localizing effects.

C. Horizontal Dipole Antenna

This antenna is in the horizontal plane, and perpendicular to the line of flight. The results apply also to an L-type antenna from the wing tip to the fuselage. In this case, the antenna does not lie in the vertical plane containing the longitudinal axis of the fuselage, and consequently α_2 in the preceding equations must be replaced by $\alpha_2 + \pi/2$. Also,

$$\beta_2 = 0.$$

Now,

$$\left.\begin{aligned} \sin\left(\alpha_2 + \frac{\pi}{2}\right) &= \cos \alpha_2 \\ \cos\left(\alpha_2 + \frac{\pi}{2}\right) &= -\sin \alpha_2 \end{aligned}\right\}. \tag{23}$$

By substitution of these values in (15), it is found that

$$\tan e = \cot \alpha_2 \tag{24}$$

and,

$$e = \frac{\pi}{2} - \alpha_2. \tag{25}$$

Thus the course error is the complement of the angle at which the airplane crosses the beacon course.

1. Localizing effect: With this antenna, a maximum signal zone occurs over the beacon. As mentioned before, this is the localizing effect

obtained in the limiting case of the inclined antenna, when $\beta_2=0$. (Fig. 3*E*.)

2. Course error: As noted immediately above, the course error is the complement of the angle at which the airplane crosses the beacon course.

3. Received voltage: From (24), it is seen that the received voltage varies as sin β_1, so that it becomes very small when the airplane is any appreciable distance from the beacon.

D. Horizontal V antenna

One variation of the horizontal antenna is the *V* antenna from wing tips to vertical fin. The performance of this type of antenna is essentially the same as that of the horizontal dipole, except that the received voltage is somewhat greater, since both horizontal components of the transmitted signal are received. The localizing effect is shown in Fig. 3*F*.

E. Symmetrical T antenna

This can be either longitudinal or transverse. The received voltage of this type of antenna is due entirely to the vertical portion, the flat top merely increasing the charging current at the top of the vertical portion, thus increasing the effective height. Consequently, the antenna will have the characteristics of its vertical or lead-in portion. If this is truly vertical, the arrangement will act the same as the simple vertical antenna, while if the lead-in portion is inclined, the characteristics will be those of the inclined antenna. The *T* antenna gives greater effective height than either type simple antenna having the same physical height because of the more uniform current distribution resulting from the additional capacity at the top of the antenna.

F. Other Antenna Systems

The general theory outlined in Section II is also applicable to other and more complicated antennas. For example, consider the inclined *V* antenna from the wing tips to the vertical fin, with the lead-in running from the junction of the two wires at the vertical fin. This is essentially a form of *T* antenna, the effects of the horizontal components perpendicular to the axis of the fuselage canceling. The antenna therefore performs as an inclined antenna having a vertical portion and also having a horizontal portion in the axis of the fuselage. The localizing effect for this type of antenna is shown in Fig. 3*G*.

Similarly, consider any antenna having a vertical lead-in and a flat-top arrangement symmetrically disposed about the vertical lead-in so that the vector sum of the horizontal projections of the individual

elements of the flat top is equal to zero. This antenna will behave in every way as a vertical antenna, but giving greater received voltage than a vertical antenna of the same physical height. It is this type of antenna which offers advantages of reduced mechanical vibration over the conventional vertical pole antenna. When the flat-top portions lie along the longitudinal axis of the fuselage, for example, the longitudinal T-antenna, the problems due to ice formation are also reduced.

IV. Experimental Results

Antennas of each of the types analyzed above were installed on the National Bureau of Standards' airplane and compared (in flight on the visual radio range beacon) with the vertical pole antenna as a standard of comparison. Relative effective heights were determined by switching the antenna under test and the pole antenna alternately to the same receiving set and recording the receiving set output voltage. Course error effects were obtained by using two receiving sets, one connected to the antenna under test and the other to the pole antenna, and comparing the course indications (as obtained on two-reed indicators connected in the output circuits of the two receiving sets), while circling the beacon at various altitudes and various distances from the beacon. The localizing effects were obtained by flying at a constant altitude on a beacon course directly over the beacon tower.

The results secured in the air were approximate only, exact quantitative measurements being very difficult because of the difficulty in measuring the geometric quantities involved. However, the results were sufficiently accurate to corroborate the theoretical analysis,

A. Relative Received Voltages

The data obtained showed that, of the antennas studied, the symmetrical T antenna of either type and the wing tip to vertical fin inclined V antenna both gave effective heights equivalent to that obtained with the vertical pole antenna without requiring nearly the same physical height as the pole antenna. For example, a symmetrical transverse T antenna having a 12-inch vertical lead-in and a flat top extending 15 feet on each side of the fuselage, parallel to and 12 inches above the wing surfaces, gave slightly better effective height than a five-foot vertical pole.

B. Course Error Effects

1. Inclined antenna ($\beta_2=20$ degrees, inclination forward). When circling the beacon at constant distance from the beacon tower, the course indications of the reed indicator operated from the receiving

set connected to the inclined antenna were in advance of the course indications of the reed indicator in the output circuit of the receiving set connected to the pole antenna. The angle of lead increased with decreasing distance from the beacon and also increased with the altitude of the airplane for a given distance from the beacon. This angle of lead is equal to the course error. The results are in accordance with (15), placing $\alpha_2 = \pi/2$, since, when circling the beacon, the line of flight is perpendicular to the line of sight. Equation (15) then resolves into (21), namely,

$$\tan e = \frac{\tan \beta_1}{\tan \beta_2}$$

where β_1 is the angle of elevation of the line connecting the airplane with the beacon (the line of sight) and β_2 the angle of inclination of the antenna (20 degrees in this case). Some idea of the magnitude of the course errors involved may be obtained from one specific case. Flying at an elevation of 3000 feet along a circle of three miles radius, the course error is 27.5 degrees.

When flying along a beacon course directly to or from the beacon, no course error was obtained. This is also in accordance with (15), since, when $\alpha_2 = 0$, equation (15) resolves into (20), namely

$$\tan e = 0.$$

2. Inclined antenna ($\beta_2 = 20$ degrees, inclination backward). The results obtained were exactly the same as for the antenna with forward inclination except that the course indications of the reed indicator operated from the receiving set connected to the inclined antenna lagged the course indications obtained on the reed indicator in the output circuit of the set connected to the pole antenna.

3. Inclined V antenna. The results secured were the same as for the inclined antenna with forward inclination.

4. Symmetrical T antenna. (Either type.) No course errors were obtained under any conditions. Both the transverse T and longitudinal T antennas were tested.

C. *Localizing Effects*

The localizing effect for each of the antennas studied was found to be very nearly the same as shown in the graphs of Fig. 3. This was perhaps the most satisfactory test of all, checking, within the limits of error, the exact theoretical analysis. At 3000 feet altitude, by changing the angle of inclination of the inclined antenna from 20 to 90 degrees, the zero signal zone could be moved from approximately one

mile from the beacon to directly over the beacon tower. When setting up the longitudinal *T* antenna, the zero signal effect directly over the beacon tower was employed as a check on the electrical symmetry of the flat-top arrangement.

V. Conclusions

For reception of signals from radio range beacons, an antenna arrangement of which the symmetrical *T* antenna is a typical example was found to give good electrical performance and to be free from trouble due to vibration. In addition, the performance of this type of antenna under conditions favoring ice formation should prove superior to that of the vertical pole antenna. To reduce the effects of mechanical vibration, this antenna employs vertical poles considerably shorter than the conventional pole antenna (10 to 18 inches instead of 5 to 6 feet). Equivalent effective height is secured through the addition of a flat top. In order to prevent directional effects, which introduce course errors, the flat top is made up of two, three, or more elements symmetrically disposed about the vertical lead-in, so that the vector sum of the horizontal projections of the individual elements is equal to zero. When but two flat-top elements are employed, the antenna takes the form of either a transverse or longitudinal *T*. The latter arrangement is preferable from the viewpoint of freedom from trouble due to ice formation. Furthermore, such a longitudinal *T* antenna has an aërodynamical resistance very much smaller than that of the transverse *T* antenna, and probably as small, if not smaller, than that of the vertical-pole antenna.

Proceedings of the Institute of Radio Engineers
Volume 20, Number 2 *February, 1932*

DISCUSSION ON "A SIMPLE METHOD OF HARMONIC ANALYSIS FOR USE IN RADIO ENGINEERING PRACTICE"*

Hans Roder

J. R. Ford:[1] Referring to Figs. 1 and 2 in the author's paper and accepting the derivation of the Fourier series, we have

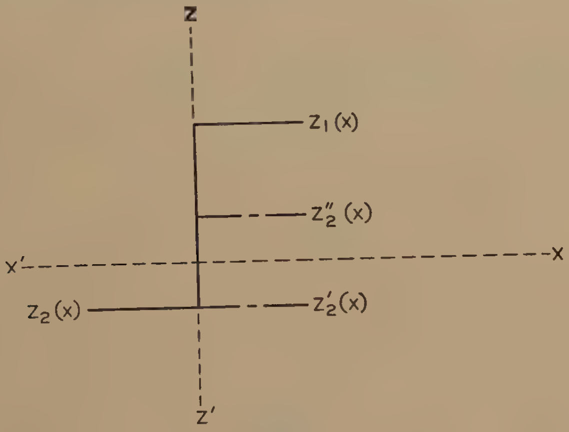

Fig. 1

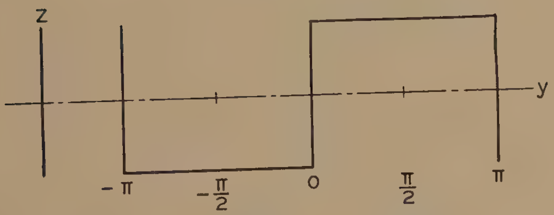

Fig. 2

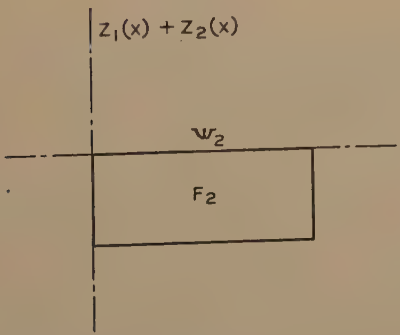

Fig. 3

* Proc. I. R. E., 19, 1481-1488, August, 1931.
[1] Narberth, Pa.

$$Z(y) = f(A \sin y) = a_0 + a_1 \sin y + b_2 \cos 2y + a_3 \sin 3y + \cdots$$

$$a_1 = \frac{1}{\pi} \int_{-\pi}^{\pi} Z(y) \sin y dy. \tag{1}$$

We plot a curve with $\cos y$ as abscissa and $Z(y)$ as ordinate and obtain as an expression for its area

$$F_A = \int_{-\pi}^{\pi} Z(y) d \cdot \cos y = -\int_{-\pi}^{\pi} Z(y) \sin y dy \text{ or } a_1 = -\frac{1}{\pi} F_A \tag{2}$$

which is substantially the same as the author's result.

However instead of using a similar method to the above in computing the magnitude of the second harmonic the author does as follows:

$$b_2 = \frac{1}{\pi} \int_{-\pi}^{\pi} Z(y) \cos 2y dy. \tag{3}$$

Referring to Fig. 2 in the author's paper we see

$$b_2 = \frac{1}{\pi} \int_{-\pi}^{0} z_2(y) \cos 2y dy + \frac{1}{\pi} \int_{0}^{\pi} z_1(y) \cos 2y dy \tag{4}$$

which the author states is equal to

$$b_2 = \frac{1}{\pi} \int_{0}^{\pi} z_1(y) \cos 2y dy - \frac{1}{\pi} \int_{0}^{\pi} z_2(y) \cos 2y dy. \tag{5}$$

so that

$$b_2 = \frac{1}{\pi} \int_{0}^{\pi} [z_1(y) - z_2(y)] \cos 2y dy. \tag{6}$$

But this is obviously untrue because $z_2(y)$ does not exist over the period $\pi > y > 0$.

We should rather find b_2 as follows:

A curve is drawn with $Z(y)$ as ordinate and $\sin 2y$ as abscissa. The area may be expressed

$$F_b = \int_{-\pi}^{\pi} Z(y) d \sin 2y = 2 \int_{-\pi}^{\pi} Z(y) \cos 2y dy. \tag{7}$$

Therefore,

$$b_2 = \frac{1}{2\pi} F_b, \quad a_3 = -\frac{1}{3\pi} F_a', \quad \text{etc.}$$

In conclusion we hope to show that in using the author's method of computing even harmonics we are reduced to an absurdity.

If the relation between $Z(x)$ and x is given graphically we may operate upon it asymmetrically resulting in the square wave form also shown below.

Computing the second harmonic (author's method) we get the area F_2 shown below.

From which (since $b_2 = (1/2\pi)2F_2$) we conclude there is some second harmonic. But in the wave form shown above only odd harmonics are present.

H. Roder:[2] In answer to the preceding discussion, it is gladly admited that the method of harmonic analysis as outlined by Mr. Ford ((2) and (7) of the above paper) can be used as well as the author's method. It is only questionable if this would be equally practicable. The author's method, namely is identical with Mr. Ford's method, except for two simplifications. According to the above discussion, we should draw curves with $Z(y)$ as ordinates and $\cos y$, $\sin 2y$, cos

[2] Radio Engineering Department, General Electric Company, Schenectady, N. Y.

$3y \cdots$ as abscissas, with y running from $-\pi$ to $+\pi$. When doing so, we obtain a curve which includes the areas:

$$\text{Area } F_{a1} \text{ for } -\pi < y < -\frac{\pi}{2}$$

$$\text{" } F_a \text{ " } -\frac{\pi}{2} < y < 0$$

$$\text{" } F_b \text{ " } 0 < y < +\frac{\pi}{2}$$

$$\text{" } F_{b1} \text{ " } +\frac{\pi}{2} < y < +\pi\,.$$

The desired Fourier coefficient is proportional to F which is the total included area

$$F = F_{a1} + F_a + F_b + F_{b1}.$$

Inspection of F shows that

$$F_a = F_{a1} \text{ and } F_b = F_{b1} \qquad (1)$$

$$F = 2(F_a + F_b).$$

F_a and F_b to have *equal* signs for odd harmonics. (2)

F_a and F_b to have *opposite* signs for even harmonics.

Thus,

$$F = 2(|F_a| + |F_b|) \text{ for odd harmonics}$$

$$F = 2(|F_a| - |F_b|) \text{ for even harmonics.}$$

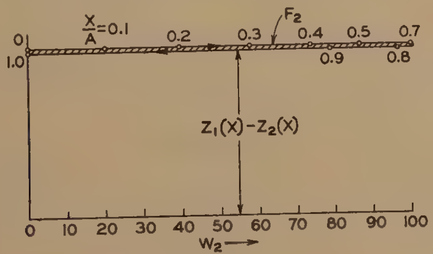

Fig. 4

The author made use of these facts by introducing the term $(z_1(x) \pm z_2(x))$ and the factor 2 (equation (1) of the paper). The magnitudes $(z_1(x) \pm z_2(x))$ are simply obtained by drawing the curves $z_2'(x)$ and $z_2''(x)$ (Fig. 2 of the paper), thus providing automatically the addition (or subtraction) of $|F_a|$ and $|F_b|$. Consequently, according to the author, the curve

$$z_1(x) \pm z_2(x) = f(w) = f_1(y)$$

is to be drawn in the interval $0 < y < +\pi/2$, while according to Ford the curve

$$Z(y) = f(w) = f_1(y)$$

is to be drawn over an interval four times as wide, namely $-\pi < y < +\pi$. This indicates less expense in time and labor for the author's method.

But also another fact is to be considered. For even harmonics, as seen above, the area F is obtained as the difference between F_a and F_b. Since F_a and F_b may

be large while their difference is small, a considerable error may result when obtaining F by separate measurement of F_a and F_b. According to the author's method, this error in drawing and measuring is reduced to a minimum, since the area $(|F_a| - |F_b|)$ can be obtained immediately by taking the vertical distance between $z_1(x)$ and $z_2''(x)$ (Fig. 2 of this paper) as ordinates for the function $(z_1(y) - z_2(y)) = f(w_2)$.

The statement made by Ford that the method does not give the correct result if applied to the square wave function in Fig. 2 cannot be maintained since it is obtained by improper application of the method. We must remember that we have to plot the function $z_1(x) - z_2(x)$ as ordinates and $w_2 = \sin 2y$ from Table II of the paper as abscissa in order to find the second harmonic. This renders a picture shown in Fig. 4. The area F_2 is obviously zero, indicating that no even harmonics are present. Thus, the graphical result is in accordance with the mathematical result.

CORRECTION

Mr. J. H. O. Harries has called the attention of Walter Schäffer and Günther Lubszynski, whose paper, "Measuring Frequency Characteristics with the Photo-Audio Generator," appeared in the July, 1931, issue of the PROCEEDINGS, to the fact that an error was made in the calculation of the screen area exposed by the perforated disk, as the formula is not

$$F = r^2 \operatorname{arc} \cos \frac{r-x}{r} - \frac{1}{2}\sqrt{r^2 - (r-x)^2} \cdot (r-x)$$

but,

$$F = r^2 \operatorname{arc} \cos \frac{r-x}{r} - \sqrt{r^2 - (r-x)^2} \cdot (r-x).$$

The factor $\frac{1}{2}$ must be omitted. This gives somewhat better approximation of the curve to the sine line.

BOOK REVIEWS

The National Physical Laboratory Report for the Year 1930. Published by His Majesty's Stationery Office, London. 295 pp. Paper binding. Price 12s 6d net.

This is a general report of the progress in the work of the following departments at the National Physical Laboratory: Physics, Electricity, Metrology, Engineering, Aërodynamics, and Metallurgy. The work of the Radio Research Board is reported by the Electricity Department. Progress is noted on experiments with the rotating loop beacon, ultra-short waves, aerials for transmission and reception and the measurement of radio-frequency current. Although the general results of the various projects are reported, it is necessary to go to other published papers to get a detailed account of the experiments. A complete list of these papers is given in the report.

*K. A. Norton

* Bureau of Standards, Washington, D. C.

This Thing Called Broadcasting, by Alfred N. Goldsmith and Austin C. Lescarboura. Published by Henry Holt and Company, New York. 362 pages, 22 illustrations, price $3.50.

This is a popular account of the development of broadcasting in the United States from the early struggles with the radiotelephone down to the present day. After the early successes of the pioneer stations broadcasting caught man's fancy and there was a rush for licenses and staking of frequency claims. At first the broadcasters were associated with the radio industry and derived their revenue from the increased sale of receiving sets and components but soon many got into broadcasting who had no such source of revenue and finally the sponsored program was devised to finance broadcasting in the United States. These developments are discussed in detail in this volume. The building of the radio program and the rôle of the announcer are interestingly described. The influence of radio on society is discussed at length in a number of chapters on music, politics, sports, women, education, the farmer, the church, and business, and the relation of each to radio. Further, broadcasting is described as a great social leveler by substituting human contact for what was previously unsound information and thus destroying the prejudices which had grown up between different social groups. This book should be interesting and instructive for both the technical and nontechnical reader as a contemporary narrative of the rise of one of the country's greatest industries, a development which has been extremely rapid and crowded with exciting incidents.

*S. S. Kirby

* Bureau of Standards, Washington, D. C.

BOOKLETS, CATALOGS, AND PAMPHLETS RECEIVED

Copies of the publications listed on this page may be obtained gratis by addressing a request to the manufacturer or publisher.

Wire-wound resistors, available with accuracy of $\frac{1}{4}$ of 1 per cent to 1 per cent are described in a four-page folder issued by the International Resistance Company of Philadelphia.

Mechanical information on three different models of variable tuning condensers is given in a six-page folder of Precise Products, Inc., 254 Mill St., Rochester, N. Y. The line of condensers illustrated are intended for manufacturers' use.

A 16-page pamphlet entitled "Weston Radio Instruments" gives technical details of meters, oscillators, test kits, and other equipment for the use of the radio experimenter, jobber, dealer, or independent service man. The pamphlet is published by the Weston Electric Instrument Corporation, of Newark, N. J.

Struthers Dunn, Inc., 139 Juniper St., Philadelphia, has recently under the name of "The Modern Lamp of Aladdin" issued a folder describing a number of their more recent products. The folder is devoted largely to a description of a light sensitive cell unit which may be used in any application which depends upon or can be made to depend upon a variation in light intensity. A number of relays, magnetically operated counters, and resistors are also described.

Bulletin No. 160 describes alternating-current portable instruments manufactured by the Roller-Smith Company, 233 Broadway, New York, N. Y. Supplement No. 2 to Bulletin No. 300 describes portable resistance meters. A metal mounting base for use with type TA, TD, or TW meters is described in Supplement No. 1 to Catalog No. 48.

"Quartz Crystals" is the title of a brochure published by the Diamond Drill Carbon Company, of 61 Park Row, New York, N. Y. dealing with the cutting and grinding of quartz crystals for use in piezo-electrically controlled vacuum tube oscillators.

Radio tube and testing equipment made by the Hickok Electrical Instrument Company, 10514 Dupont Ave., Cleveland, O. is described in an eight-page folder. The equipment described is primarily of interest to servicemen and radio dealers.

Proceedings of the Institute of Radio Engineers
Volume 20, Number 2 *February, 1932*

RADIO ABSTRACTS AND REFERENCES

THIS is prepared monthly by the Bureau of Standards,* and is intended to cover the more important papers of interest to the professional radio engineer which have recently appeared in periodicals, books, etc. The number at the left of each reference classifies the reference by subject, in accordance with the "Classification of Radio Subjects: An extension of the Dewey Decimal System," Bureau of Standards Circular No. 385, which appeared in full on pp. 1433–56 of the August, 1930, issue of the PROCEEDINGS of the Institute of Radio Engineers.

The articles listed are not obtainable from the Government or the Institute of Radio Engineers, except when publications thereof. The various periodicals can be secured from their publishers and can be consulted at large public libraries.

R000. RADIO

R007.9 Second meeting of the International Technical Consulting Committee on Radio Communication, Copenhagen, 1931. PROC. I.R.E., **19**, 2219–2249; December, 1931.

"The twenty-one opinions which were unanimously approved at the 1931 C.C.I.R. meeting are given together with fourteen questions for consideration at the third meeting to be held in Lisbon, Portugal, at a time to be specified later. A list of the members of the U. S. delegation is included."

R100. RADIO PRINCIPLES

R113 Phenomena accompanying radio transmission. (A translation of an address by Marchese Marconi given on September 11, 1930 at Treno before the Italian Society for the Progress of Sciences.) *Marconi Review*, 1–8; September-October, 1931.

Some phenomena which accompany radio transmissions are treated in a somewhat summary and elementary manner.

R113 K. Försterling and H. Lassen. Die Ionisation der Atmosphäre und die Ausbreitung der kurzen elektrischen Wellen (10–100 m) über die Erde. (The ionization of the atmosphere and the propagation of short electro-magnetic waves along the earth's surface.) *Zeit. für tech. Physik*, **12**, 453–469, No. 10; 1931; 502–527, No. 11; 1931.

A comprehensive discussion and mathematical study of the problems confronting short wave radio communication.

R113.2 J. Asakura. On the results of the continuous record of atmospherics. *Reports of Radio Researches and Works in Japan*, **1**, 113–126; August, 1931.

A sketch of the recorder and the statistical results obtained over a period of three years are described.

R113.5 A. Kimpara. Correlation of atmospherics with thunder-storms. *Reports of Radio Researches and Works in Japan*, **1**, 126–150; August, 1931.

The method of measurement is given. The author discusses the relation between atmospherics and thunderstorms, typhoons, discontinuous surfaces, and distant thunderstorms. Several graphs and tables of data are presented.

* This list compiled by Mr. A. H. Hodge, Mr. W. H. Orton, and Miss E. M. Zandonini.

R113.61 T. R. Gilliland and G. W. Kenrick. Preliminary note on an automatic recorder giving a continuous height record of the Kennelly-Heaviside layer. Research Paper 373. *Bureau of Standards Journal of Research*, **7**, 783–790; November, 1931.

"A description is given of a preliminary installation of a recorder which gives a continuous automatic record of the virtual height of the Kenelly-Heaviside layer. The group retardation method of Breit and Tuve is employed with modifications which permit a continuous record to be made. Suggestions are made for improvements which might be incorporated in a permanent installation."

R113.61 H. Rakshit. On an estimation of the height of the Heaviside layer in Bengal. *Phil. Mag.*, **12**, 897–907; November, 1931.

The natural fading method used to obtain an estimate of the height of the Kennelly-Heaviside layer apparatus is described.

R133 H. E. Hollman and T. Schultes. Raumakustische Kippschwingungen. (Acoustic relaxation oscillations.) *Elek. Nach-technik*, **8**, 494–502; November, 1931.

An arrangement for generating relaxation oscillations is described, wherein the usual condenser is replaced by an inclosed air space containing a microphone and loudspeaker. Equations for determining frequency and the acoustic properties of the inclosure are given.

R133 F. B. Llewellyn. Constant frequency oscillators. Proc. I.R.E., **19**, 2063–2094; December, 1931.

"The manner in which the frequency of vacuum tube oscillators depend upon the operating voltages is discussed. The theory of the dependence is derived and is shown to indicate methods of causing the frequency to be independent of the operating voltages. These methods are applied in detail to the more commonly used oscillator circuits. Experimental data are cited which show the degree of frequency stability which may be expected as a result of application of the methods outlined in the theory. The appendix contains an analysis of the conditions under which the performance of an oscillator may be represented by the use of linear circuit equations."

R133 J. B. Dow. A recent development in vacuum tube oscillator circuits. Proc. I.R.E., **19**, 2095–2108; December, 1931.

"A constant frequency oscillator, which depends for its operation upon the use of electron coupling between the oscillation generating portion of the circuit and the work circuit is described. This form of coupling is employed to isolate the work circuit from the frequency determining portion of the system. The oscillator is of the two-anode type (UX-865) and it is shown that by suitable choice of anode voltages, compensating effects may be obtained whereby changes in generator voltage may be made to have a negligible effect upon the frequency of oscillation."

R133 C. K. Jen. A new treatment of electron tube oscillators with feed-back coupling. Proc. I.R.E., **19**, 2109–2144; December, 1931.

This article gives a mathematical discussion of electron tube oscillators which have a feed-back coupling between grid and plate outside of the tube. Several graphs and experimental data are included.

R133 W. Reichardt. Entartungserscheinungen am Röhrensender. (Degenerate phenomena in vacuum tube oscillators.) *Elek. Nach.-technik*, **8**, 502–512; November, 1931.

Factors governing generated wave form in vacuum tube oscillators are discussed. Conditions for very irregular waves are illustrated graphically.

R133 H. E. Hollman. Über symmetrische Kippschwingungen und ihre Synchronisierung. (Symmetrical relaxation oscillations and their synchronization.) *Elek. Nach.-technik*, **8**, 449–457; October, 1931.

A symmetrical vacuum tube arrangement for generating relaxation oscillations is analyzed and its mode of operation explained. Practical applications are suggested.

R134 W. Greenwood and S. J. Preston. Quality detectors. *Wireless Engineer and Experimental Wireless*, **8**, 648–658; December, 1931.

"The object of this article is to indicate how the various practical methods of rectification can be used to give the best quality, according to conclusions which have been arrived at as a result of work carried out in this subject during the past few years."

R140 A. Boyajian. Mathematical analysis of non-linear circuits. *General Electric Review*, **34**, 531; September, 1931; 745; December, 1931.

"The departure of actual voltages, currents, and magnetic fluxes from theoretical sine waves greatly complicates a complete mathematical analysis in modern design. In this article curves representing various phenomena are resolved into "regions" which are in turn resolved into two easily handled components, thus affecting a simple solution of problems that otherwise would present much difficulty."

R140 C. L. Suits. Non-linear circuits for relay applications. *Electrical Engineering*, **50**, 963; December, 1931.

Study and analysis of non-linear circuits reveal characteristics which mark them as capable of being adapted to a variety of special uses. A simple series circuit of this type is especially well adapted for relay applications, and obviates some of the undesirable features of mechanical relays.

R142 Decoupling. *Wireless World and Radio Review*, **24**, 550–552; November 11, 1931.

In modern sets it is necessary to use decoupling. Graphs are given which should be of service in selecting circuit constants.

R144 K. Tani. Radiation resistance of complex antennas. *Reports of Radio Researches and Works in Japan*, **1**, 185–197; August, 1931.

A mathematical treatment of the radiation resistance of complex antennas is given. Tables of radiation resistance are included.

R146 Y. Kusunose. Elimination of harmonics in valve transmitters. *Reports of Radio Researches and Works in Japan*, **1**, 151–155; August, 1931.

A brief description is given of the present state of technique concerning the supression of harmonics radiated from vacuum tube transmitters. A method of eliminating the strongest harmonic is suggested.

R148 H. Roder. Amplitude, phase, and frequency modulation. PROC. I.R.E., **19**, 2145–2176; December, 1931.

"This paper presents a comparative theoretical study of amplitude, phase, and frequency modulation." A few cases of "false phase modulation" are treated.

R161.4 A. L. M. Sowerby. The output stage and the loudspeaker. *Wireless World and Radio Review*, **24**, 585–589; November 18, 1931.

This article explains how the greatest undistorted output is obtained from the last stage.

R165 D. A. Woliver. The inductor dynamic loudspeaker. (Merits revealed by measured performance data.) *Wireless World and Radio Review*, **24**, 579–582; November 18, 1931.

As well as giving precise details of the behavior of the inductor this article includes useful information concerning the working conditions.

R200. Radio Measurements and Standardization

R201.7 F. T. Brewer. Cathode-ray oscillograph timing axis. *Electronics*, **3**, 222–223; December, 1931.

An arrangement consisting of a condenser which is charged from the plate current of a 224 radio tube and discharged through a voltage regulator tube is used to provide a time axis for a cathode-ray oscillograph. The frequency range is 25 to 5000 cycles per second.

R201.7 H. E. Hollman. Die Aufnahme nichtperiodischer Vorgänge mit dem Kathodenstrahloszillographen. (Recording non-periodic phenomena with the cathode ray oscillograph.) *Archiv für Elek.*, **25**, 689–694; October, 1931.

Description of method and practical applications are given.

R210 A. J. Makower and W. Makower. A new method of measuring frequencies. *Jour. Sci. Instr.*, **8**, 286–287; September, 1931.

The arrangement under electrostatic forces of lycopodium powder in ridges on a moving slate enables alternating-current frequencies to be measured by counting the ridges formed.

R220 ×R230 F. M. Colebrook. The double beat method of frequency adjustment. *The Wireless Engineer and Experimental Wireless*, **8**, 639–646; December, 1931.

Applications of the double beat method of frequency adjustment to the measurement of capacity and inductance are made. Circuits, equations, and graphs are included.

R220 K. Kuhlmann. Messcondensator mit einer von exakt Null linear ansteigenden Kapazität. (A measuring condenser with linear capacity variation down to exactly zero.) *Archiv für Elek.*, **25**, 666–668; October, 1931.

Details of construction and methods of capacity measurement are given.

R223 ×R145.5 M. J. O. Strutt. Dielektrische Eigenschaften verschiedener Gläser in Abhängigkeit der Frequenz und der Temperatur. (Dielectric properties of different types of glass in relation to frequency and temperature.) *Archiv für Elek.*, **25**, 715–722; October, 1931.

The dielectric constants and loss angles of five different glasses were measured as functions of temperature and frequency and an empirical relationship set up which is discussed in the light of the dipole theory.

R240 ×621.313.7 W. Wolman and H. Kaden. Die Anwendung des Trockengleichrichters in der Tonfrequenz-Messtechnik. (The application of dry rectifiers to audio-frequency measuring methods.) *Zeit. für tech. Physik*, **12**, 470–482, No. 10, 1931.

Circuit arrangements for utilizing dry rectifiers in the measurement of current, voltage, and damping at audio frequencies are given. It is shown that temperature variations may be compensated and that accurate, sensitive measurements are possible.

R241.5 L. O'Bryan. The portable double bridge. *General Electric Review*, **34**, 752; December, 1931.

A portable double bridge designed for making low resistance measurements is described.

R254 A. H. Cooper and G. P. Smith. A direct-reading modulation meter. *The Wireless Engineer and Experimental Wireless*, **8**, 647; December, 1931.

The circuit and constants of the meter are given.

R254 ×537.7 M. Grutzmacher. Die Fourieranalyse-modulierter Hochfrequenz. (The Fourier analysis of modulated high-frequency waves.) *Elek. Nach.-technik*, **8**, 476–480; November, 1931.

Direct determination of the amplitude and frequency of component waves, in a modulated high-frequency wave, is accomplished by the apparatus and procedure described.

R254 W. Runge. Die Untersuchung amplituden-und frequenz-modulierter Sender. (Measurements on amplitude and frequency modulated transmitters). *Elek. Zeit.*, **52**, 1322–1323; October, 22, 1931.

The author describes apparatus for determining the presence and degree of phase or frequency modulation, or distortion in the output of a vacuum tube transmitter. The method is due to Gritzmacher and consists essentially of harmonic analysis.

R254 A. Heilman. Ein stroboskopisches Verfahren zur Messung von Frequenz-und Phasenmodulation. (A stroboscopic method for measuring frequency and phase modulation.) *Elek. Nach.-technik*, **8**, 469–476; November, 1931.

A new method is described for measuring the degree of phase and frequency modulation present in an amplitude modulated transmitter. A Braun tube is employed.

R300. Radio Apparatus and Equipment

R330 New electron tubes—Facts and rumors. *Electronics*, **3**, 216–217; December, 1931.

Descriptions of a new short-wave tube for television, a two-in-one power tube and the radio-frequency pentode are given. The preliminary rating and characteristics of the new radio-frequency pentode are listed.

R331 Cobalt alloy filament. *Electronics*, **3**, 232; December, 1931.

A cobalt alloy filament announced by the DeForest Radio Company is said to overcome handicaps experienced by the '30, '31, and '32 types of tubes.

R334 W. T. Cocking and W. I. G. Page. The advantages of the variable-mu valve—Simplified circuit design and elimination of noise. *Wireless World and Radio Review*, **24**, 546–548; November 11, 1931.

The merits and characteristics of the new tube are presented. It is predicted that this tube will replace the screen-grid tube in many of its present uses. A circuit diagram of a radio-frequency amplifier which employs two variable-mu stages is given.

R335 D. F. Schmit. More power to midgets—The new power pentode tube. *Radio News*, **13**, 19; July, 1931.

Data on power pentode tubes are given.

R335 J. R. Nelson. Pentode tubes used as triodes. *Electronics*, **3**, 226–227; December, 1931.

Various characteristic curves of the pentode are plotted. The operation of the pentode is compared to that of the triode.

R337 P. Schwerin. Replacing the type '80 rectifier with a mercury vapor tube. *Radio Craft*, **3**, 401; January, 1932.

A discussion of the advantages and the method of installing the new mercury vapor rectifier are given.

R339 H. J. Reich. A self-stopping d.c. thyratron circuit. *Electronics*, **3**, 240; December, 1931.

A circuit by which a condenser is charged through a high resistance and discharged through a thyratron is explained.

R339 B. S. Havens. Industry adopts the electron tube. *General Electric*
×535.38 *Review*, **34**, 714; December, 1931.

Photo-electric and thyratron tube controls that have already found wide application and that offer fruitful possibilities in the industrial plant are described in this article.

R355.6 Y. Kusunose and S. Ishikawa. Frequency stabilization of radio transmitters. *Reports of Radio Researches and Works in Japan*, **1**, 157–183; August, 1931.

Methods of stabilizing a master oscillator are given.

R355.65 O. M. Hovgaard. A new oscillator for broadcast frequencies. *Bell Lab. Record*, **10**, 106–110; December, 1931.

A well-constructed piezo oscillator which has small frequency variations is described.

R355.9 D. Lewis. How to build your own beat frequency oscillator. *Radio News*, **13**, 580–582; January, 1932.

Complete design data for a beat frequency oscillator are given.

R355.9 W. Fucks. Ein einfacher Stossgenerator für einmalige und periodische Vorgänge. (A simple impulse generator for single and periodic impulses.) *Archiv für Elek.*, **25**, 723–744; November, 1931.

An impulse generator which operates on the dynatron principle is described. Its utility in supplying a time axis for the oscillograph and in field intensity measuring apparatus is explained.

R356.2 N. Pomeranz. Power supply for the Radio News transmitter. *Radio News*, **13**, 571; January, 1932.

The power supply for a one-tube transmitter is described. The advantages of the 566 mercury vapor rectifier are discussed. Some operating hints are also included.

R360 N. R. Bligh and E. D. Whitehead. Controlling volume with the variable-mu valve. *Wireless World and Radio Review*, **24**, 606–607; November 25, 1931.

A method of controlling volume by keeping the screen voltage constant under varying conditions is given.

R361 C. H. Johnson. A new "super" for circuit experimenters. *Radio News* **13**, 578–579; January, 1932.

Description of a JSW-4, an all wave set using homemade coils and offering considerable flexibility as far as other parts are concerned is given.

R365.2 ×R145.5 H. Vogt. Der tönende Kondensator. (The sounding condenser.) *Elek. Zeit.*, **52**, 1402–1407; November 12, 1931.

After describing several types of sound reproducers, the author discusses principles of operation and construction of the electrostatic loud speaker.

R385.5 A. H. Reeves. Étude d'un microphone pour la radio-diffusion (Study of a microphone for radiotelephony.) *L'Onde Electrique*, **10**, 458–470; October, 1931.

A study is made of a microphone relatively free from noise and capable of reproducing uniformly the frequencies from 50–10,000 cycles per second, and possessing a linear characteristic for all amplitudes compatable with ordinary amplifier tubes.

R388 J. B. Johnson. The cathode-ray oscillograph. *Jour. Frank. Inst.*, **212**, 687–717; December, 1931.

The author describes the operation, structure, and uses of cathode-ray oscillographs.

R400. Radio Communication Systems

R410 C. R. Burch. On asymmetric telegraphic spectra. Proc. I.R.E., **19**, 2191–2218; December, 1931.

It is shown that single side band Morse transmission, if practicable, would relieve the present long-wave spectral congestion. Methods are developed whereby the wave shape of the single side band signals can be visualized when the original message envelope is given, and it is shown that the prolonged transmission of true single side band signals would in general necessitate the radiation of infinite amplitudes. Wave forms which evade this difficulty are determined. The production and reception of asymmetric side band waves is discussed.

R412 Ship-to-shore telephone installation on the S. S. Empress of Britain. *Marconi Review*, 18–25; September-October, 1931.

The difficulties associated with long-distance ship-to-shore telephony, and the methods used to overcome them are outlined in the description of the wireless telephone installation on the "Empress of Britain." A special feature of the equipment is the ability to speak from the ordinary cabin telephones which are connected to the wireless telephone through the ship's manual switchboard so that telephone service to and from the ship is as convenient as the normal shore telephone service.

R413.1 ×R365.3 A. Bailey and T. A. McCann. Application of printing telegraph to long-wave radio circuits. Proc. I.R.E., **19**, 2177–2190; December, 1931.

"This paper describes certain arrangements which have been used for start-stop printing telegraph operation over a transatlantic long-wave radio channel and also describes results obtained from certain tests of long-wave teletypewriter transmission from Rocky Point, L.I., to Rochester, N. Y. A prediction of year-round results is obtainable by correlation of these test data with year-round noise measurement data taken at Houlton, Maine, in connection with transatlantic telephone service."

R430 J. M. Duguid. Reduction of radio interference from telephone power plants. *Bell Lab. Record*, **10**, 124–126; December, 1931.

The method used by the Western Electric Company to reduce radio interference is described.

R430 Radio interference. *Radio News*, **13**, 560–561; January, 1932.

This article points out some methods of checking up on interference produced in homes and of eliminating this interference.

R430 K. Heinrich. Über eine Möglichkeit, Rundfunkstörungen zu unterdrücken, die durch elektrische Schaltwerke entstehen. (The elimination of broadcast interference caused by sign flashers.) *Elek. Zeit.*, **52**, 1358–1359; October 29, 1931.

It is shown that a small half-watt lamp connected across the flasher switch will eliminate serious interference.

R500. Applications of Radio

R510 Coastal and harbor wireless services. *Marconi Review*, 9–17; September-October, 1931.

Equipments are briefly described and mention is made of the more recent applications of the wireless art to marine navigation.

R526 Serre. La radiogoniométrie appliquée aux lignes aériennes. (Radiogoniometry applied to air lines.) *L'Onde Electrique*, **10**, 425–457; October, 1931; 513–520; November, 1931.

The explanation of some ideas on the subject of the protection of the radiogoniometer against nocturnal disturbances.

R526.12 New aircraft beacon, visual type course indicator for Croydon Aerodrome. *Electrician*, **107**, 756; November 29, 1931.

A visual type course indicator is to be installed at London airport, Croydon.

R533 Electronic equipment in train control. *Electronics*, **3**, 218–220; December, 1931.

Automatic vacuum tube apparatus is used to continuously signal to the cab the condition of the road ahead. Apparatus and methods of operation are discussed.

R583 J. Calcaterra. How to build a home television receiver. *Radio News*, **13**, 37–40; July, 1931.

Constructional details of such a receiving set are given.

R800. Nonradio Subjects

534 N. Andrejew. Drei einfache Methoden der technischen Akustik. (Three simple methods of acoustic measurement.) *Elek. Nach.-technik*, **8**, 488–494; November, 1931.

Three simple methods of determining the frequency response of vibrating membranes are described.

535.38 G. F. Metcalf and A. J. King. A new selenium tube. *Electronics*, **3**, 234–235; December, 1931.

The characteristics and uses of the selenium cell are given.

535.38 M. C. Teves. Über hochempfindliche vacuum-photoelektrische Zellen. (Highly sensitive vacuum-photo-cells.) *Zeit. für tech. Physik*, **12**, 556–558; No. 11; 1931.

A vacuum enclosed, oxide-coated caesium photocell is described which has a sensitivity of 65 micro-amperes per lumen.

537.65 V. Petrzilka. Über den Zusammenhang zwischen den optischen und piezoelektrischen Eigenschaften der schwingenden Quarzplatten. (On the relation between the optical and piezo-electric properties of vibrating quartz crystals.) *Annalen der Physik*, **11**, 623–632, No. 5; 1931.

An optical method of observing vibrating quartz crystals under polarized light is shown to be valuable in determining their piezo-electric properties.

621.313.7 New metal rectifier—The Westinghouse H.T.8 unit for voltage-doubling circuits. *Wireless World and Radio Review*, **24**, 560–561; November 11, 1931.

Data and instructions for using metal rectifiers are included.

621.314.6 Air-cored chokes for tone correction. *Wireless World and Radio Review*, **24**, 553; November 11, 1931.

Constructional data for chokes of high inductance.

621.382 E. H. B. Bartelink and G. H. Bast. Die Übertragung von Telegraphenzeichen. (Telegraph transmission.) *Elek. Nach.-technik*, **8**, 480–488; November, 1931.

A contribution to the theory and design of telegraph circuits.

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